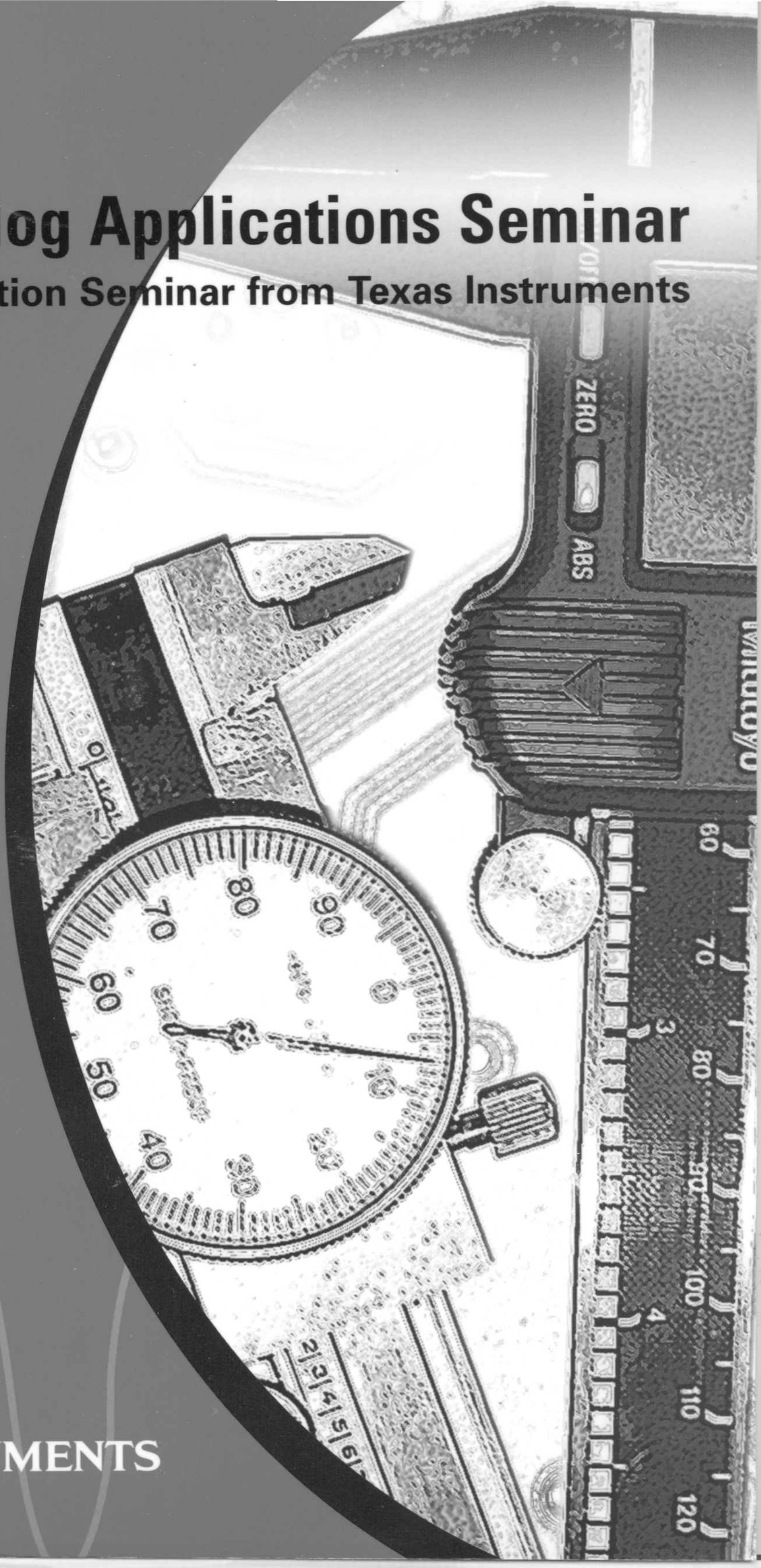
Precision Analog Applications Seminar

A User's Application Seminar from Texas Instruments



TEXAS INSTRUMENTS

AUBREY.

Precision Analog Applications Seminar

Precision Analog Applications Seminar

Texas Instruments

delta A/D

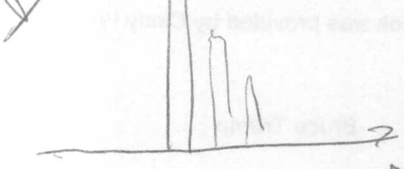
RMS noise ± σ (S.D.) 67% time value won't chge.

peak −2.6 bits worse ± 3.3σ

99% of time value won't chge.

Linearity
end point
on best fit

endpoint worse
performance



noise
histogram

gaussian distribution → white noise

↑ show then DNL
but width problems.

Texas Instruments

Introduction

Texas Instruments is pleased to present the Precision Analog Applications Seminar to our customers. This detailed technical seminar focuses on applications where physical parameters - temperature, light, weight - must be measured with precision. The focus is on commonly encountered circuit applications with an eye on the tricky issues that arise in all analog designs.

The first application touches on measuring many different physical phenomena from a remote location. Sensor characteristics are discussed, and signal conditioning for these sensor outputs to make them suitable for conversion to digital signals is detailed.

Second, a detailed look at thermocouples and how they work is offered. This common temperature-sensing element is widely used but often misunderstood by circuit designers. Removing the mysteries surrounding thermocouples and showing how their signals can be properly conditioned for use in measurement systems provides designers with real-world advice they can apply directly.

Data acquisition systems which have a number of inputs from different sensors or signal sources give rise to several approaches to tackling this problem. Should you use separate converters for each signal, or multiplex the signals into one? What issues arise from using delta-sigma converters in multiplexed systems? These questions are addressed and guidelines proposed for each of these scenarios.

Finally, sensors used in a bridge arrangement are common - for measuring temperature, pressure, and weight. The last application focuses on these bridge measurement systems and discusses interfacing with converters, what signal conditioning and filtering should be used, and ends with a detailed look at a reference design from TI which implements a high-precision weigh scale.

Throughout these application topics, sidebar discussions will be introduced, to look in more detail at useful tools and techniques to improve your designs. Simulating circuits with SPICE, designing active filters, considerations for driving successive-approximation (SAR) analog-to-digital converters are all discussed. Some important concerns with how printed circuit board layout affects precision are also considered.

This seminar was written by the applications staff of Texas Instruments. The authors are Russell Anderson, Michael Ashton, Rick Downs, and Thomas Kuehl.

Additional support in producing this book was provided by Cindy Williams and Charles Wray.

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Texas Instruments	Texas Instruments

Precision Analog Applications Seminar

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Precision Analog Applications Seminar

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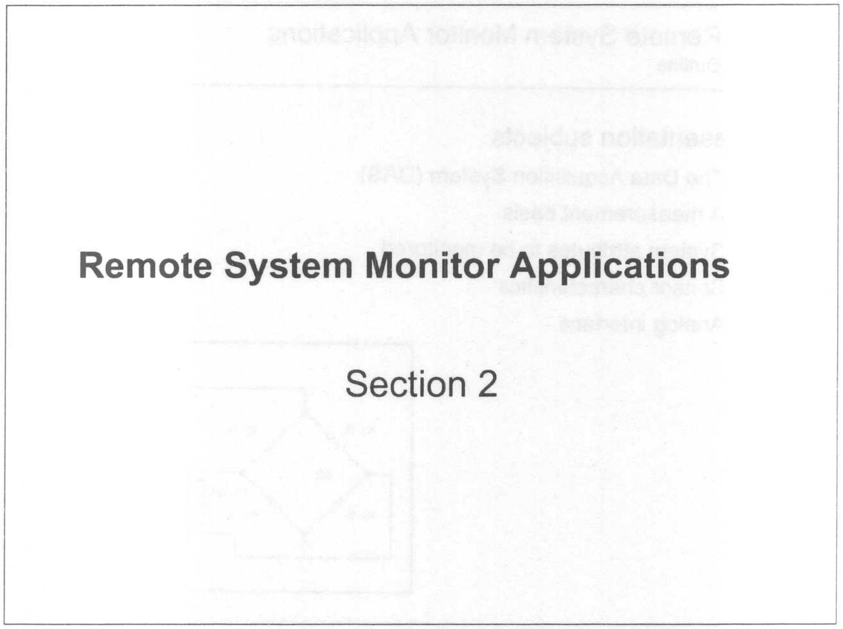
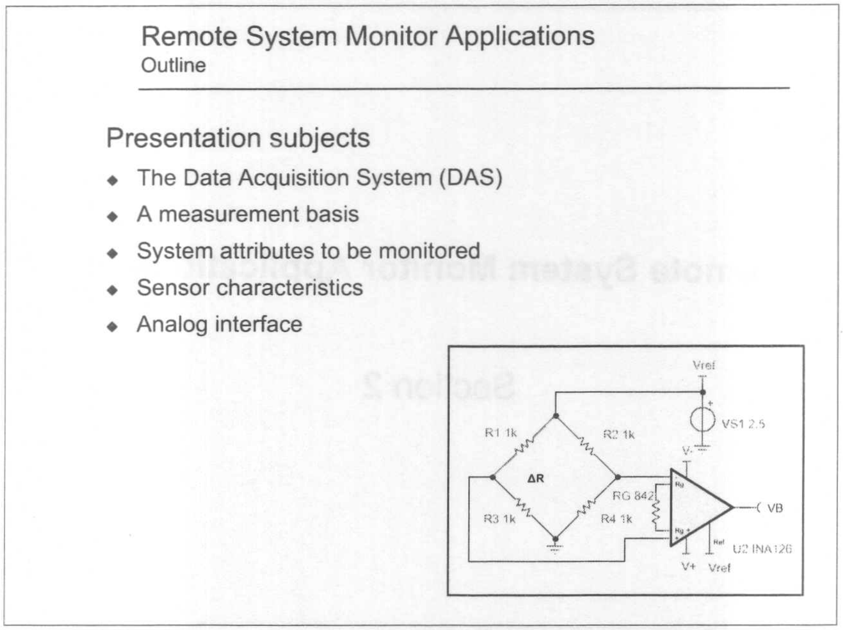


Remote System Monitor Applications

Section 2

Welcome to the 2005 Precision Analog Applications Seminar. This portion of the seminar will focus on Remote System Monitor applications. We'll take a look at sensors and analog circuits that can be applied in monitoring the operating conditions of a remotely located system.



Remote System Monitor Applications

Outline

Presentation subjects

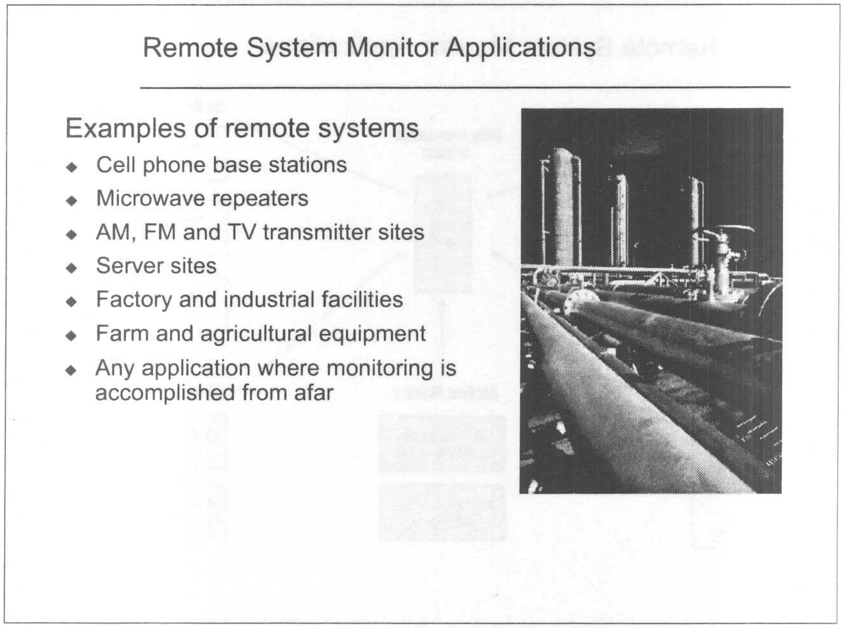
- The Data Acquisition System (DAS)
- A measurement basis
- System attributes to be monitored
- Sensor characteristics
- Analog interface

Vref
VS1 2.5
R1 1k
R2 1k
V-
ΔR
RG 842
R3 1k
R4 1k
Rg
Rg
Ref
VB
U2 INA126
V+
Vref

This seminar will focus on the task of remotely measuring and monitoring a system's operating parameters. An integrated data acquisition system (DAS) IC will be included in the discussion, as it serves as the conduit for collecting and converting the signals from the various sensor channels.

The measurement bridge circuit that serves as the fundamental circuit for many sensors will be reviewed. Several types of sensors that work as an integral part of the measurement bridge will be discussed. The sensor response characteristics will be discussed in some detail.

The bridge response to a stimulus is then conditioned by an analog interface circuit where it may be amplified, filtered, level shifted, etc, before being applied to a DAS input. Various circuits for accomplishing these tasks will be presented.



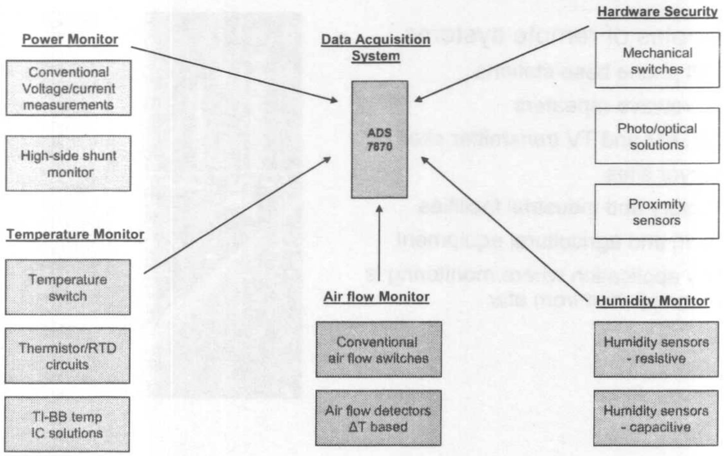
Remote System Monitor Applications

Examples of remote systems

- Cell phone base stations
- Microwave repeaters
- AM, FM and TV transmitter sites
- Server sites
- Factory and industrial facilities
- Farm and agricultural equipment
- Any application where monitoring is accomplished from afar

A remote system may be any system that requires monitoring from a different location. Connection to the remote system may be made via cable, telephone lines, fiber optic lines, or wireless links.

Remote System Monitor Applications



Power Monitor

Conventional Voltage/current measurements

High-side shunt monitor

Temperature Monitor

Temperature switch

Thermistor/RTD circuits

TI-BB temp IC solutions

Data Acquisition System

ADS 7870

Air flow Monitor

Conventional air flow switches

Air flow detectors ΔT based

Hardware Security

Mechanical switches

Photo/optical solutions

Proximity sensors

Humidity Monitor

Humidity sensors - resistive

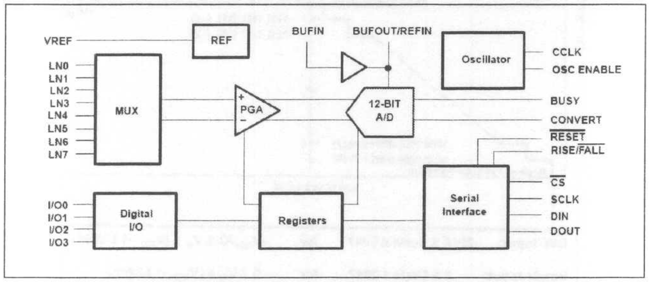
Humidity sensors - capacitive

Here are some examples of different system attributes that may be monitored in a remote installation. An ADS7870 data acquisition system IC (DAS) collects and converts the information from the sensors to a digital serial format. Although some sensors may have sufficient output to directly drive the analog input scale of the DAS, many won't. Most often an analog interface is required to amplify and condition the sensor outputs before they are applied to the DAS inputs.

Remote System Monitor Applications

ADS7870 Data Acquisition System

- 12-bit, 52k-samples/sec DAS with A/D, MUX, PGA and reference
- PGA gains 1, 2, 4, 5, 8, 10, 16 and 20V/V
- Programmable inputs - Up to 4-ch differential or 8-ch single ended
- Selectable internal reference of 1.15V, 2.048V, or 2.5V (or ext input)
- 2.7V to 5.5V single supply operation
- 4-bit digital I/O serial interface - SPI™, QSPI™, Microwire™ and 8051-family protocols, without glue logic



ADS7870 Data Acquisition System Diagram

Shown here is the ADS7870 DAS block diagram. The 4 differential (8 single-ended) multiplexer inputs allow for a corresponding number of analog channels to be simultaneously monitored.

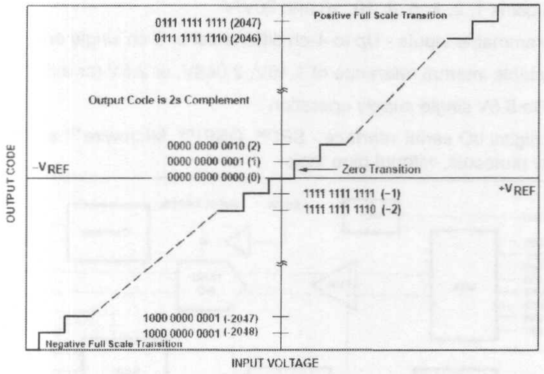
The ADS7870's PGA features selectable gain levels that are electronically set. Thus, a wide range of input signals can be accommodated. The internal reference, selectable at 1.15V, 2.048V and 2.5V, provides a convenient reference voltage that can be used to bias the reference pin of instrumentation amplifiers.

For many remote monitoring applications the 12-bit A/D resolution will provide adequate resolution.

Remote System Monitor Applications

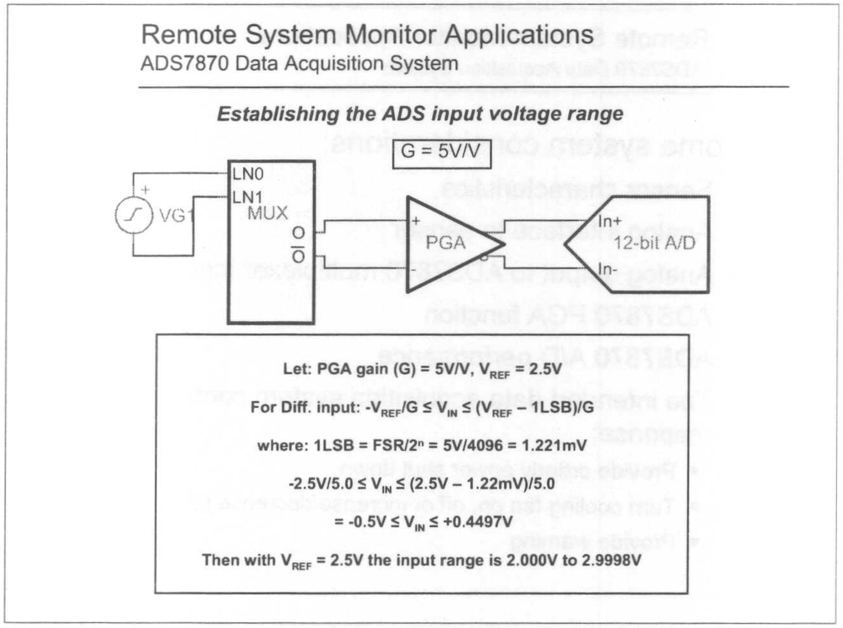
ADS7870 Data Acquisition System

ADS7870 output codes vs. input voltage



Diff. input:	**$-2048 \le \text{Code} \le 2047$**	***for***	**$-V_{REF}/G \le V_{IN} \le (V_{REF}$ -1 LSB)/G**
Single input:	**$0 \le \text{Code} \le 2047$**	***for***	**$0 \le V_{IN} \le (V_{REF}$ -1 LSB)/G**

The ADS7870's internal 12-bit, A/D converter provides 2^{12}, or 4096 distinct output codes, for an input range of –FS (full scale) to +FS. Depending on the input mode, single-ended or differential, the code range is 0 to 2047 or -2048 to 2047, respectively. The converter's output code is a 2's complementary binary format.



Remote System Monitor Applications

ADS7870 Data Acquisition System

Establishing the ADS input voltage range

Let: PGA gain (G) = 5V/V, V_{REF} = 2.5V

For Diff. input: $-V_{REF}/G \le V_{IN} \le (V_{REF} - 1LSB)/G$

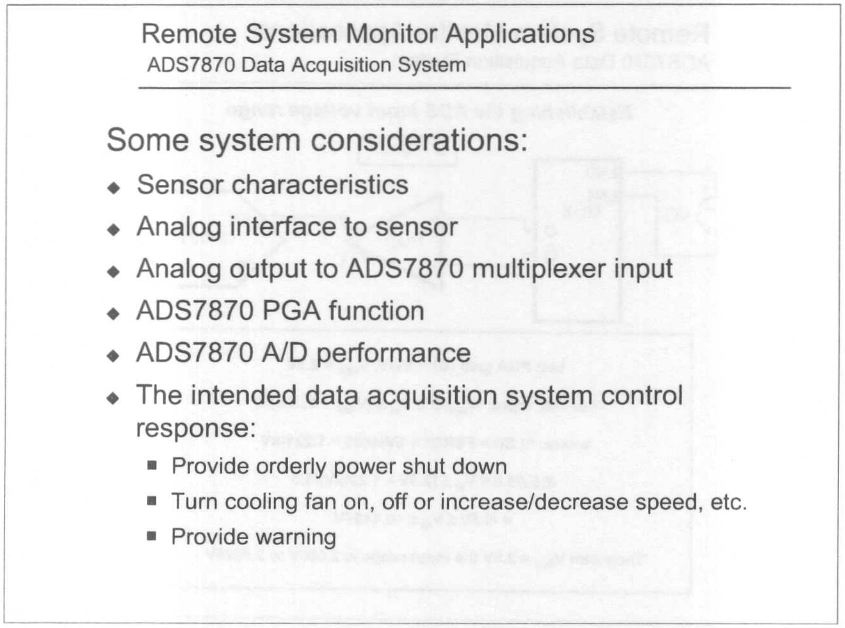
where: $1LSB = FSR/2^n = 5V/4096 = 1.221mV$

$-2.5V/5.0 \le V_{IN} \le (2.5V - 1.22mV)/5.0$

$= -0.5V \le V_{IN} \le +0.4497V$

Then with V_{REF} = 2.5V the input range is 2.000V to 2.9998V

Whenever possible use the maximum input range to capitalize on the A/D converter's full-scale resolution. The slide shows how to establish the A/D converter's input voltage, taking into account the selected reference voltage and PGA gain.



Remote System Monitor Applications

ADS7870 Data Acquisition System

Some system considerations:

- Sensor characteristics
- Analog interface to sensor
- Analog output to ADS7870 multiplexer input
- ADS7870 PGA function
- ADS7870 A/D performance
- The intended data acquisition system control response:
    - Provide orderly power shut down
    - Turn cooling fan on, off or increase/decrease speed, etc.
    - Provide warning

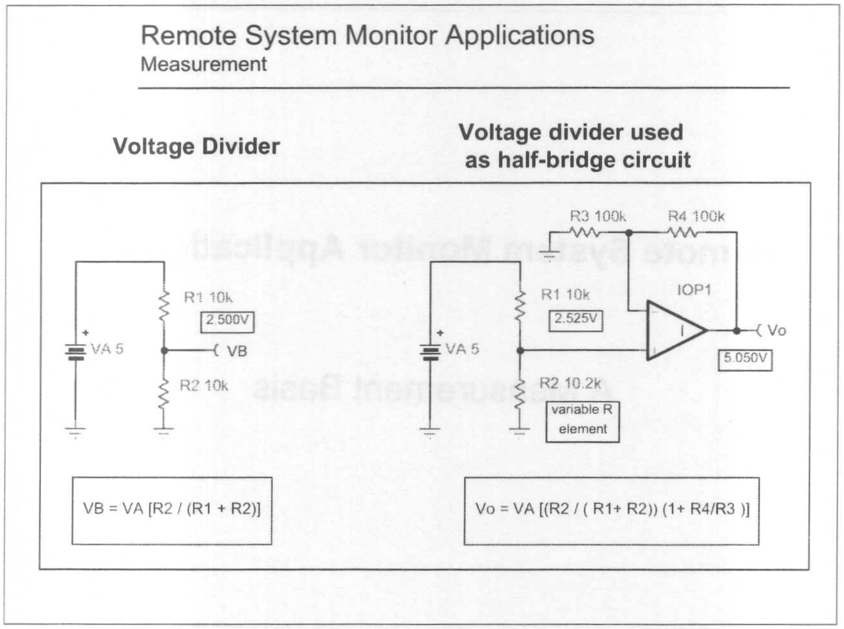
The various sensors and transducers have widely different electrical characteristics. This ultimately dictates the analog interface circuit design and DAS set up for that measurement channel. The DAS should be set up to use the full A/D input range. This improves the system signal-to-noise ratio.

Should the DAS output code indicate an out-of-range or fault condition a decision needs to be what the system response will be.

Remote System Monitor Applications

A Measurement Basis

It is helpful to establish a measurement basis before proceeding with the sensors and analog interface circuits.



Remote System Monitor Applications

Measurement

Voltage Divider

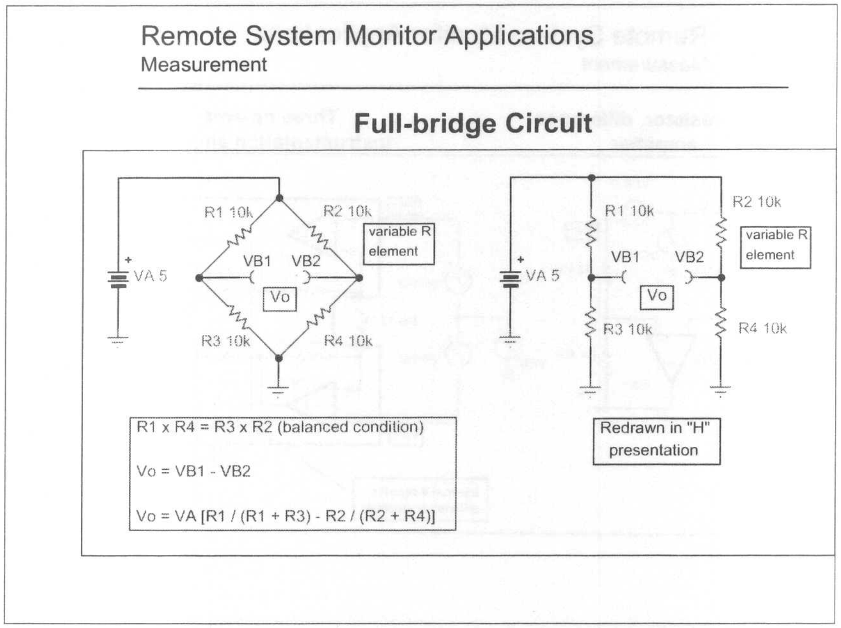
Voltage divider used as half-bridge circuit

$VB = VA\,[R2 / (R1 + R2)]$

$Vo = VA\,[(R2 / (R1 + R2))\,(1 + R4/R3)]$

The basis for many sensor circuits is the simple voltage divider; where one of the resistors in the divider is a resistive sensor. The voltage at the divider union, VB, changes in response to a change in the sensor resistance. This simple circuit is often referred to as a half bridge. Note that a load connected to VB will reduce the normal unloaded value measured and must be taken into consideration.

The half bridge output can be buffered with an operational amplifier which also can be configured for a voltage gain. The op-amp should be selected to minimize the loading on the divider *i.e.* have low input bias current.



Remote System Monitor Applications

Measurement

Full-bridge Circuit

R1 x R4 = R3 x R2 (balanced condition)

Vo = VB1 - VB2

Vo = VA [R1 / (R1 + R3) - R2 / (R2 + R4)]

Redrawn in "H" presentation

If we take the simple voltage divider, or half-bridge, and place 2 of them back-to back a full-bridge is created. When all resistors are equal, the output voltage at each half is identical. Changing one or more resistors unbalances the bridge resulting in a differential voltage change (Vo).

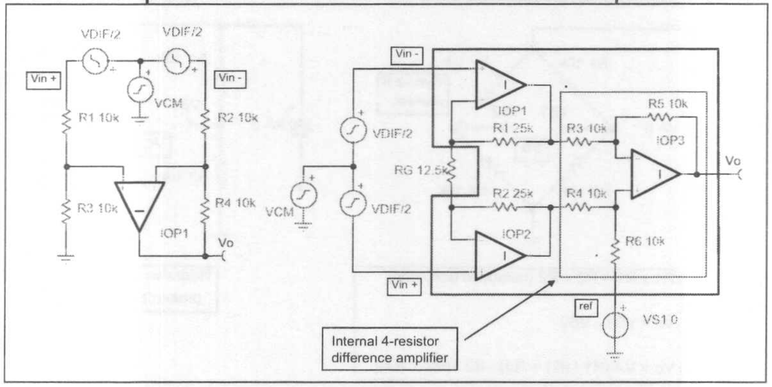
The full-bridge is sometimes referred to as a measurement bridge. And although it has a similar appearance to a Wheatstone bridge, the two perform different functions.

The Wheatstone bridge incorporates a current meter at points VB1 and VB2, and is used to indicated when a balanced condition is achieved. At that point the voltage at VB1 and VB2 are equal resulting in no current flow through the meter path. Unknown resistance or reactance can be inferred by knowing the resistances or reactances in the other bridge legs.

Remote System Monitor Applications

Measurement

Four-resistor, difference amplifier **Three op-amp, instrumentation amplifier**



Internal 4-resistor difference amplifier

A difference amplifier is formed when the 4-resistor bridge is connected to an op-amp as shown. A common mode voltage (Vcm) applied to the two op-amp inputs, will be cancelled by the equal and opposite responses of the two inputs. If all resistors are exactly equal in value and the common mode rejection is very high the output voltage will be nearly zero. A differential voltage applied to the difference amplifier inputs appears out of phase at the amplifier and is thus amplified.

The difference amplifier has an output voltage relationship:

Differential $Vo_{Dif} = (Vin+) - (Vin-)$
Common-mode $Vo_{CM} = [(Vin+) + (Vin-)] / 2$

The VDif and Vcm source arrangement shown in the diagrams is an analysis model that conveniently allows both the differential and common-mode sources to be simultaneously applied. A split differential source is not a practical circuit in reality.

By buffering the difference amp with a differential input gain stage a three op-amp, instrumentation amplifier is created. The instrumentation amplifier serves as the interface between the bridge and any subsequent signal conditioning circuits. It provides excellent common-mode rejection and high gain accuracy.

Remote System Monitor Applications

Temperature Monitoring

Remote System Monitor Applications

Temperature control and monitoring – on/off solution

ThermTROL

YOUR PARTNER IN PRECISE THERMAL CONTROL AND PROTECTION

Our Products

1 NT SERIES 1/2 DISC THERMOSTATS

Thermtrol gives you both High Quality and Packaging Options.

Features

- **Mechanical Solution – bimetal switch**
- **Fixed temperature thermostats**
- **On-off operation from 0°C to 200°C**
- **AC power switching**
- **Low voltage DC power switch available**
- **Manufactured by Thermtrol, Airpax, Selco, Thermodisc, Matsuo and others**

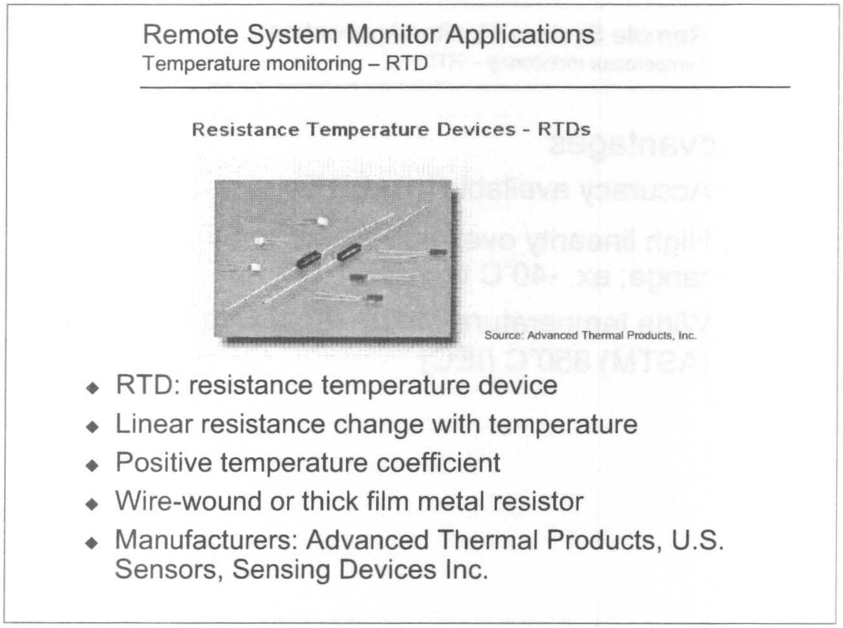
1. The thermal switch is an easily implemented, on/off control.

2. Tolerance is different for open and close.

Automatic reset type

Temp range C	open ± C	close ± C
0 -93	3.0	4.0
94-121	3.5	4.5
122-149	4.0	5.5
150-204	5.0	7.0

3. Suppliers include Thermtrol, Airpax, Selco, Termodisc, Matsuo and others.



Remote System Monitor Applications

Temperature monitoring – RTD

Resistance Temperature Devices - RTDs

Source: Advanced Thermal Products, Inc.

- RTD: resistance temperature device
- Linear resistance change with temperature
- Positive temperature coefficient
- Wire-wound or thick film metal resistor
- Manufacturers: Advanced Thermal Products, U.S. Sensors, Sensing Devices Inc.

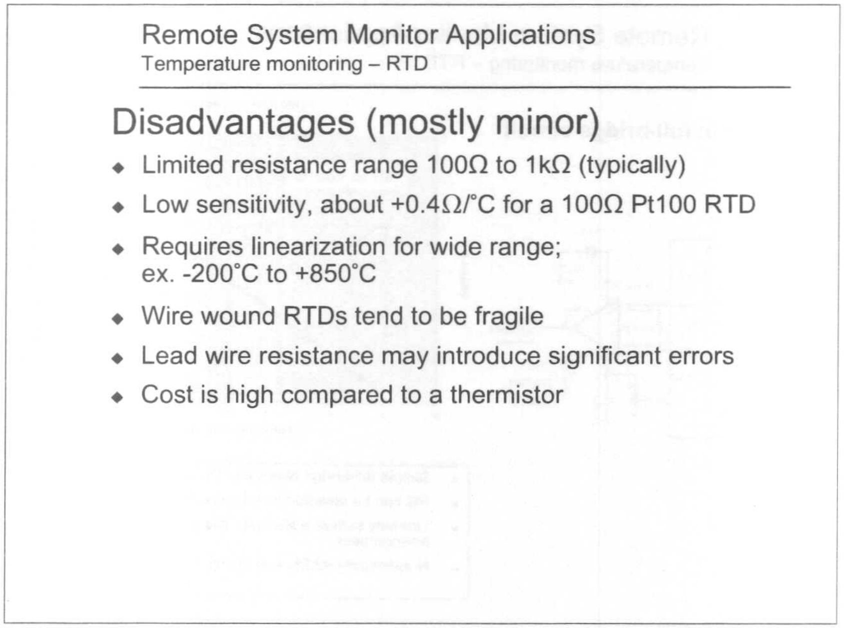
The resistance temperature device, RTD, provides accurate, moderate cost temperature sensing. An important characteristic of the RTD is its near linear resistance change across temperature, $\Delta R/\Delta T$. Note, however, that the best linearity is achieved within portions of the full operating temperature range.

Remote System Monitor Applications

Temperature monitoring – RTD

Advantages

- Accuracy available to ±0.1˚C
- High linearity over limited temperature range; ex. -40˚C to +85˚C
- Wide temperature range: -250˚C to 600˚C (ASTM) 850˚C (IEC)



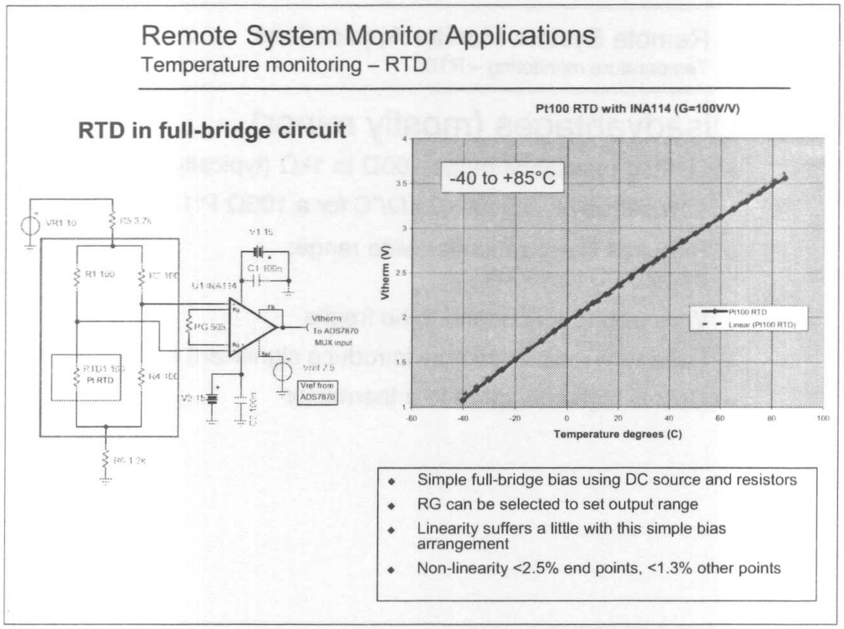
Remote System Monitor Applications

Temperature monitoring – RTD

Disadvantages (mostly minor)

- Limited resistance range 100Ω to 1kΩ (typically)
- Low sensitivity, about +0.4Ω/°C for a 100Ω Pt100 RTD
- Requires linearization for wide range; ex. -200°C to +850°C
- Wire wound RTDs tend to be fragile
- Lead wire resistance may introduce significant errors
- Cost is high compared to a thermistor

Since the RTD is a low resistance device long lead wires can add resistance in the sensors path and cause errors. Three and 4 wire solutions exist and can be used to correct for this additional resistance component.



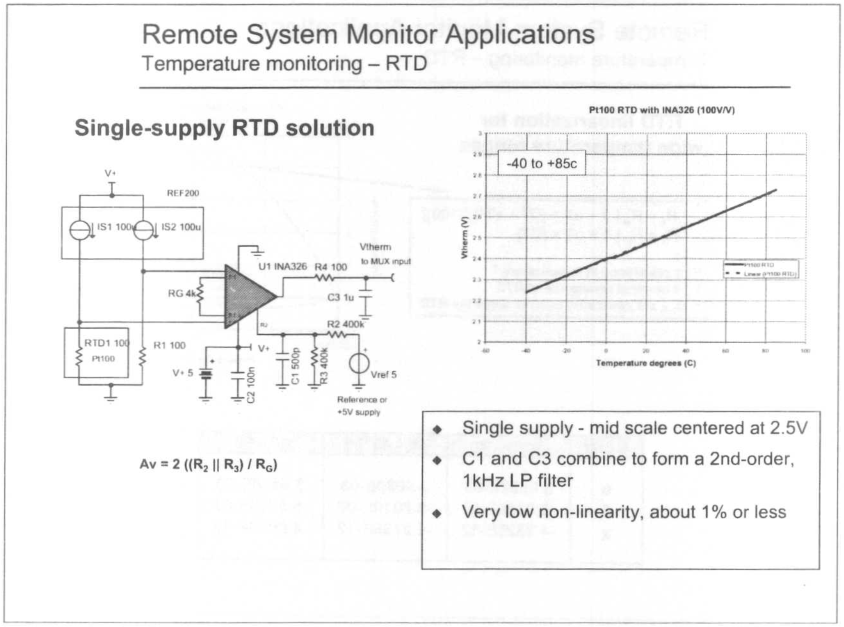
Remote System Monitor Applications

Temperature monitoring – RTD

RTD in full-bridge circuit

- Simple full-bridge bias using DC source and resistors
- RG can be selected to set output range
- Linearity suffers a little with this simple bias arrangement
- Non-linearity <2.5% end points, <1.3% other points

The schematic shows an RTD connected in a full-bridge circuit. The resistors above and below the bridge were selected such that the voltage at both INA114 inputs is 2.5V when the bridge is balanced. This matches the reference voltage which was selected as the mid scale voltage for the ADS7870 A/D converter. Reasonably good linearity performance is achieved from -40 to +85°C.



Remote System Monitor Applications

Temperature monitoring – RTD

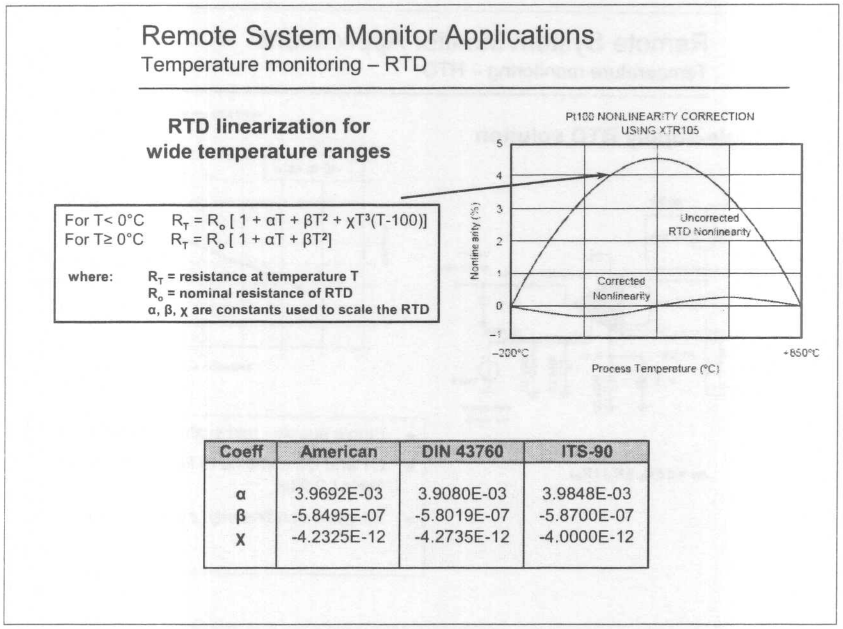
Single-supply RTD solution

$Av = 2\,((R_2 \,||\, R_3) / R_G)$

- Single supply - mid scale centered at 2.5V
- C1 and C3 combine to form a 2nd-order, 1kHz LP filter
- Very low non-linearity, about 1% or less

This schematic illustrates a more precise approach to biasing the RTD. The upper resistors in the full-bridge have been replaced by two equal, low current, current sources. This is easily accomplished with a REF200. The bridge bias current is low, 100µA, so the RTD dissipates little power, and an improvement in linearity is had over the previous circuit.

An INA326 is employed as a single-supply, bridge amplifier. The INA326 circuit can be configured to include a 2nd-order, low-pass filter, which can be helpful in reducing the noise response. The output of this sensor circuit can be further amplified using the ADS7870's internal PGA.



Remote System Monitor Applications

Temperature monitoring – RTD

RTD linearization for wide temperature ranges

For $T < 0°C$ $R_T = R_o [1 + \alpha T + \beta T^2 + \chi T^3(T-100)]$
For $T \geq 0°C$ $R_T = R_o [1 + \alpha T + \beta T^2]$

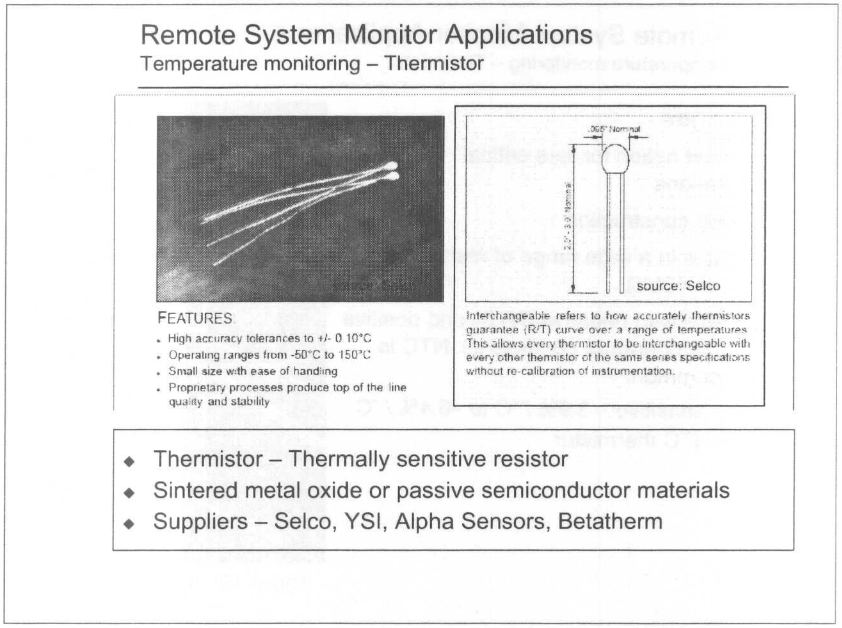
where: R_T **= resistance at temperature T**
R_o **= nominal resistance of RTD**
α, β, χ **are constants used to scale the RTD**

Coeff	American	DIN 43760	ITS-90
α	3.9692E-03	3.9080E-03	3.9848E-03
β	-5.8495E-07	-5.8019E-07	-5.8700E-07
χ	-4.2325E-12	-4.2735E-12	-4.0000E-12

Although the intended application of this seminar is a remote system monitor where temperature ranges are likely to be quite limited, there may be applications where wide temperature ranges must be monitored with high accuracy.

The RTD linearity has a predictable quadratic characteristic with a peak deviation approaching 5%. This quadratic curvature is shown in the accompanying diagram. Since it is predictable it is possible to reduce the error by applying linearization techniques.

The second curve in the diagram shows the reduction in the linearity error after linearization is applied. This particular RTD linearization technique is applicable to an XTR105, 4-20mA, 2-wire transmitter. The technique of summing a secondary current into the RTD across temperature reduces the error to that shown in the second curve.



Remote System Monitor Applications

Temperature monitoring – Thermistor

source: Selco
.065" Nominal
2.0" - 3.0" Nominal

source: Selco

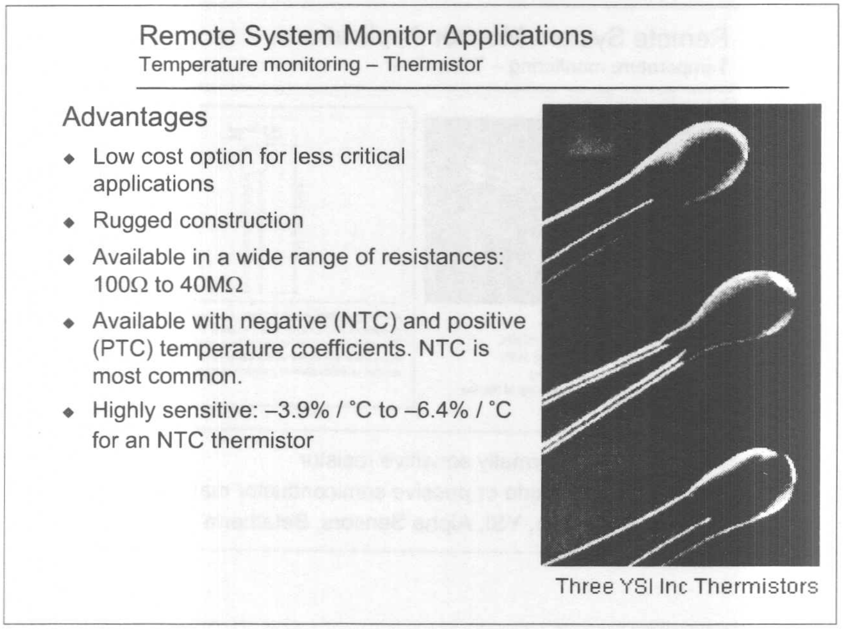
FEATURES

- High accuracy tolerances to +/- 0.10°C
- Operating ranges from -50°C to 150°C
- Small size with ease of handling
- Proprietary processes produce top of the line quality and stability

Interchangeable refers to how accurately thermistors guarantee (R/T) curve over a range of temperatures. This allows every thermistor to be interchangeable with every other thermistor of the same series specifications without re-calibration of instrumentation.

- Thermistor – Thermally sensitive resistor
- Sintered metal oxide or passive semiconductor materials
- Suppliers – Selco, YSI, Alpha Sensors, Betatherm

The thermistor serves well as an economical temperature sensor for less critical applications. It's available in a wide range of resistance values and with tolerance values to 0.1%. Note, however, that this is the tolerance specified at a specific temperature and not over a wide temperature range.



Remote System Monitor Applications

Temperature monitoring – Thermistor

Advantages

- Low cost option for less critical applications
- Rugged construction
- Available in a wide range of resistances: 100Ω to 40MΩ
- Available with negative (NTC) and positive (PTC) temperature coefficients. NTC is most common.
- Highly sensitive: –3.9% / ˚C to –6.4% / ˚C for an NTC thermistor

Three YSI Inc Thermistors

The NTC thermistor is by far the most common and is often used as a temperature sensor. The PTC thermistor is often employed as a temperature sensitive element within a circuit. It is often used to alter the bias of a circuit in response to temperature.

Remote System Monitor Applications

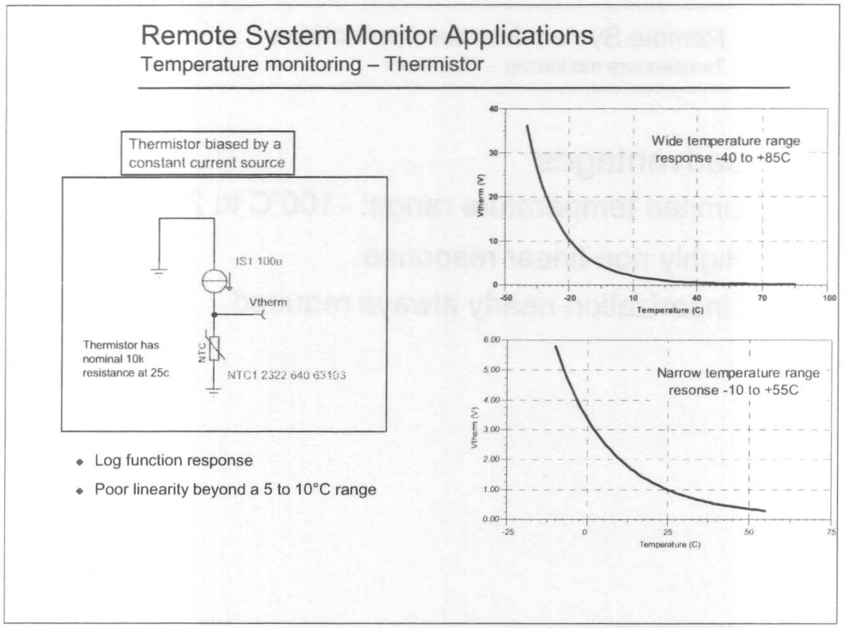
Temperature monitoring – Thermistor

Disadvantages

- Limited temperature range: –100°C to 200°C
- Highly non-linear response
- Linearization nearly always required

Considering the attractiveness of thermistor's low cost, often less than $1.00, the downside isn't too bad. The primary drawback of the device is the poor linearity performance if the application calls for monitoring a wide temperature range.

For many limited remote monitoring applications, the thermistor's usable range and linearity is completely adequate - once simple linearization techniques are applied.



Remote System Monitor Applications

Temperature monitoring – Thermistor

Thermistor biased by a constant current source

Thermistor has nominal 10k resistance at 25c

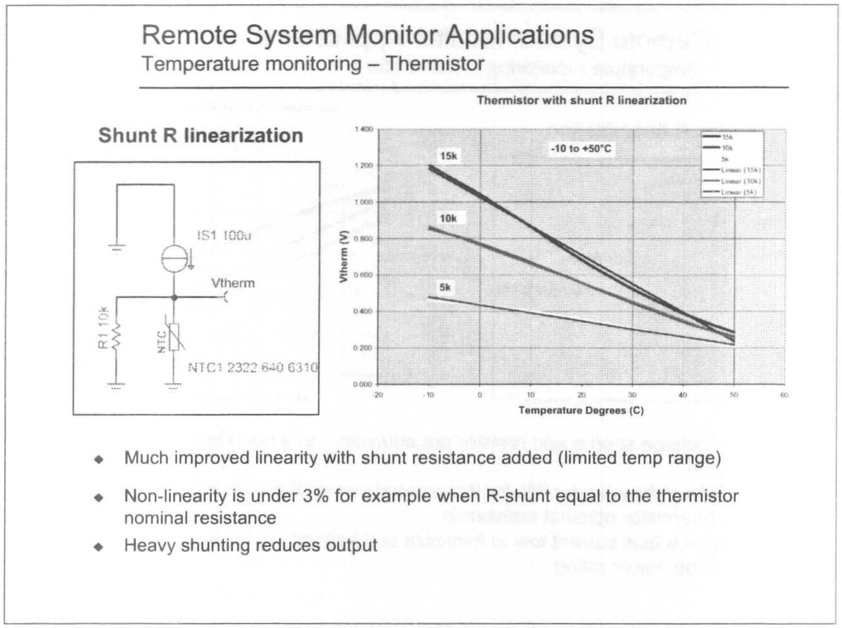
- Log function response
- Poor linearity beyond a 5 to 10°C range

Wide temperature range response -40 to +85C

Narrow temperature range resonse -10 to +55C

IS1 100u
Vtherm
NTC
NTC1 2322 640 63103
40
30
20
10
0
Vtherm (V)
-50
-20
10
40
70
100
Temperature (C)
6.00
5.00
4.00
3.00
2.00
1.00
0.00
Vtherm (V)
-25
0
25
50
75
Temperature (C)

These plots depict the thermistor's response to temperature over a narrow and then wider temperature range. Thermistor biasing is by a simple current source. As is, the linearity suffers with temperature spans greater than 5 to 10°C.



Remote System Monitor Applications

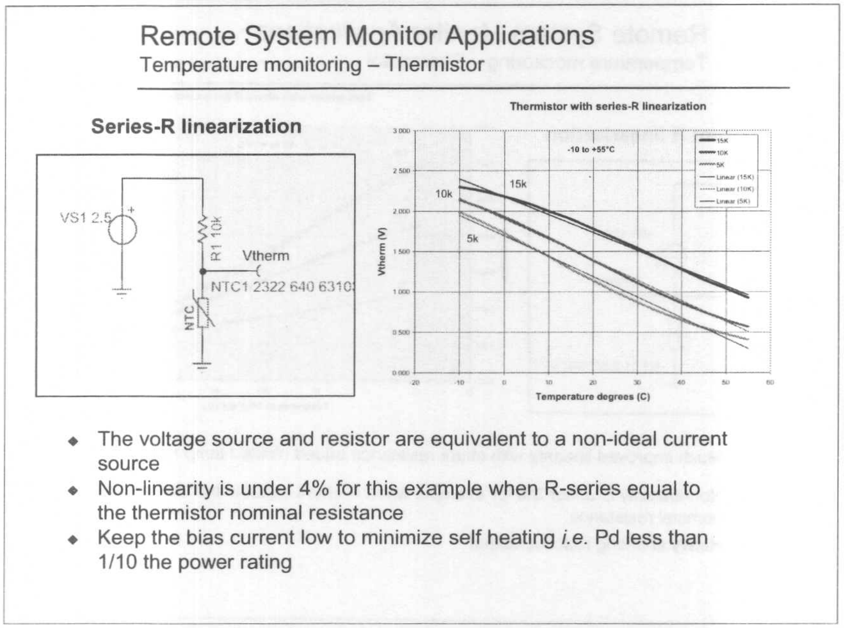
Temperature monitoring – Thermistor

Shunt R linearization

IS1 100u
Vtherm
R1 10k
NTC
NTC1 2322 640 6310
Thermistor with shunt R linearization
-10 to +50°C
15k
10k
5k
Linear (15k)
Linear (10k)
Linear (5k)
Vtherm (V)
1.400
1.200
1.000
0.800
0.600
0.400
0.200
0.000
-20
-10
0
10
20
30
40
50
60
Temperature Degrees (C)

- Much improved linearity with shunt resistance added (limited temp range)
- Non-linearity is under 3% for example when R-shunt equal to the thermistor nominal resistance
- Heavy shunting reduces output

Shunting the thermistor with a resistor drastically improves the linearity across temperature. Note that the output voltage range is reduced as the shunt resistor is reduced in value. The best compromise between linearity and output voltage is achieved when the shunt resistor is equal to the thermistor resistance.



Remote System Monitor Applications

Temperature monitoring – Thermistor

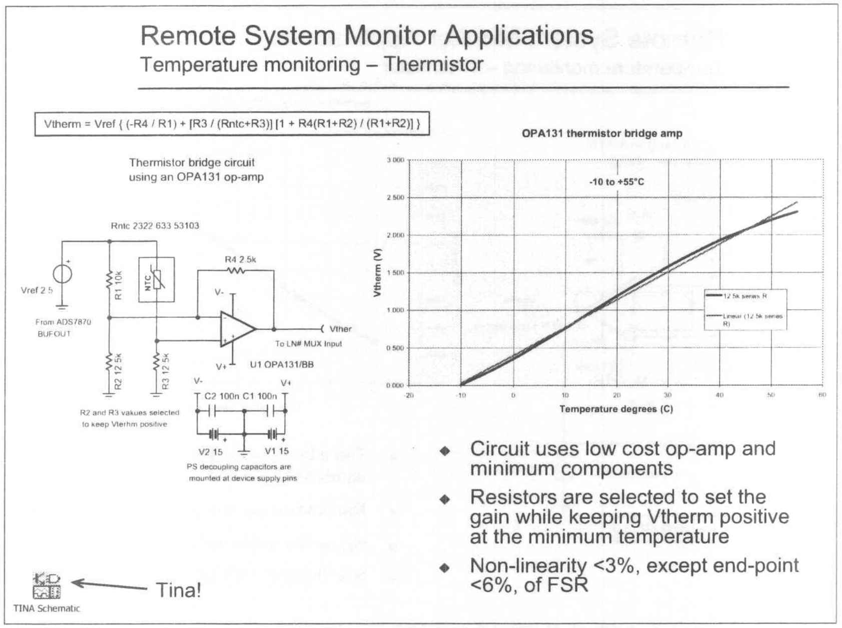
Series-R linearization

VS1 2.5
+
R1 10k
Vtherm
NTC1 2322 640 6310
NTC
Thermistor with series-R linearization
-10 to +55°C
15K
10K
5K
Linear (15K)
Linear (10K)
Linear (5K)
15k
10k
5k
Vtherm (V)
3 000
2 500
2 000
1 500
1 000
0 500
0 000
-20
-10
0
10
20
30
40
50
60
Temperature degrees (C)

- The voltage source and resistor are equivalent to a non-ideal current source
- Non-linearity is under 4% for this example when R-series equal to the thermistor nominal resistance
- Keep the bias current low to minimize self heating *i.e.* Pd less than 1/10 the power rating

A thermistor may also be biased by a voltage source. A series resistor is added to establish the current through the thermistor. The voltage source and series resistor create the equivalent of a simple current source.

Like the shunt linearization circuit the series implementation provides improved thermistor linearity performance across temperature. Again best results are had when the series resistor is made equal in value to that of the thermistor.



Remote System Monitor Applications

Temperature monitoring – Thermistor

Vtherm = Vref { (-R4 / R1) + [R3 / (Rntc+R3)] [1 + R4(R1+R2) / (R1+R2)] }

Thermistor bridge circuit using an OPA131 op-amp

Rntc 2322 633 53103
Vref 2.5
From ADS7870 BUFOUT
R1 10k
NTC
R4 2.5k
V-
V+
Vther
To LN# MUX Input
U1 OPA131/BB
R2 12.5k
R3 12.5k
C2 100n
C1 100n
V2 15
V1 15

R2 and R3 vakues selected to keep Vterhm positive

PS decoupling capacitors are mounted at device supply pins

OPA131 thermistor bridge amp

-10 to +55°C
Vtherm (V)
3.000
2.500
2.000
1.500
1.000
0.500
0.000
-20
-10
0
10
20
30
40
50
60
Temperature degrees (C)
12.5k series R
Linear (12.5k series R)

- Circuit uses low cost op-amp and minimum components
- Resistors are selected to set the gain while keeping Vtherm positive at the minimum temperature
- Non-linearity <3%, except end-point <6%, of FSR

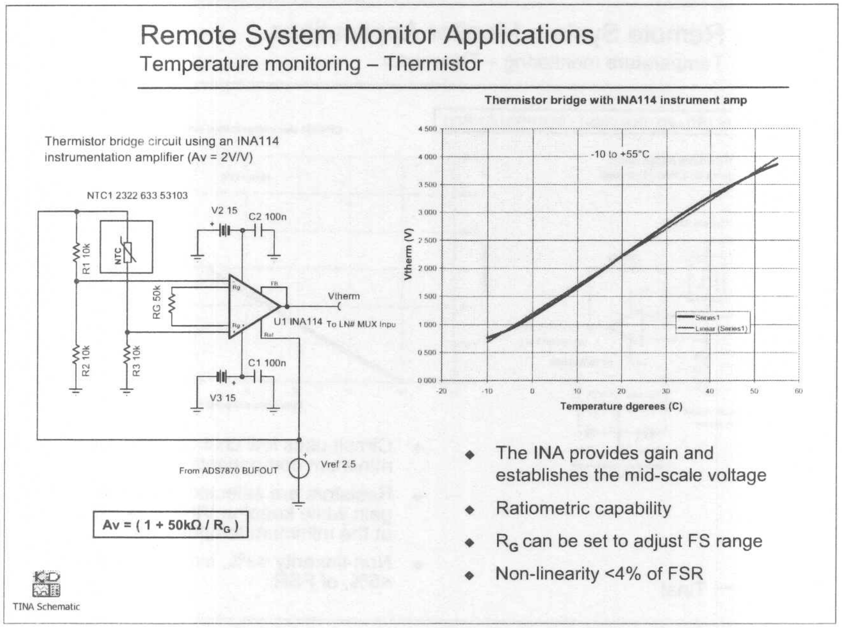
Tina!

TINA Schematic

This is a low cost, full-bridge circuit for a thermistor temperature sensor. The OPA131 that follows the bridge amplifies the voltage difference at the centers of the 2 legs in the bridge.

The circuit is a little tricky to set because the resistor values must be selected to set the gain and mid-scale voltage. The plot shows a linearity response that is adequate for many non-critical applications.

Notice the Tina Schematic symbol on the lower left corner of the slide. Tina is a powerful simulation tool that was used to analyze many of the circuits in this presentation. It is ideal for analyzing the sensor circuit performance across temperature.



Remote System Monitor Applications

Temperature monitoring – Thermistor

Thermistor bridge circuit using an INA114 instrumentation amplifier (Av = 2V/V)

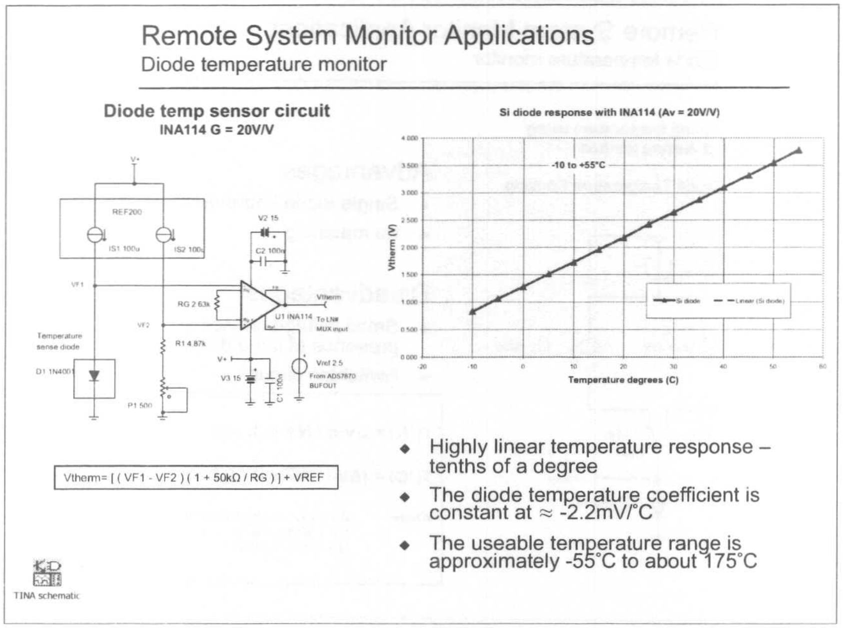
NTC1 2322 633 53103
V2 15
C2 100n
R1 10k
NTC
RG 50k
Rg
FB
Vtherm
U1 INA114
To LN# MUX Inpu
Ref
R2 10k
R3 10k
C1 100n
V3 15
From ADS7870 BUFOUT
Vref 2.5
TINA Schematic
Thermistor bridge with INA114 instrument amp
Vtherm (V)
4.500
4.000
3.500
3.000
2.500
2.000
1.500
1.000
0.500
0.000
-20
-10
0
10
20
30
40
50
60
Temperature dgerees (C)
-10 to +55°C
Series1
Linear (Series1)

$$Av = (1 + 50k\Omega / R_G)$$

- The INA provides gain and establishes the mid-scale voltage
- Ratiometric capability
- R_G can be set to adjust FS range
- Non-linearity <4% of FSR

The INA114 instrumentation amplifier is an ideal interface between the bridge and the DAS system:

1. The INA voltage gain can be set as needed. Keep in mind that the DAS has the internal PGA as well.
2. A reference voltage can be applied to INA's Ref pin, to match the DAS mid-scale voltage.
3. The reference voltage can also be used to bias the bridge allowing the bridge bias to respond in a ratiometric fashion. Since the reference is established in the DAS, the internal A/D, the INA114 reference voltage, and the bridge bias voltage all track.


V+
REF200
IS1 100u
IS2 100u
V2 15
C2 100n
VF1
VF2
RG 2.63k
R1 4.87k
Vtherm
U1 INA114
To LN#
MUX input
Temperature
sense diode
D1 1N4001
P1 500
V+
V3 15
C1 100n
Vref 2.5
From ADS7870
BUFOUT
Si diode response with INA114 (Av = 20V/V)
-10 to +55°C
Vtherm (V)
4.000
3.500
3.000
2.500
2.000
1.500
1.000
0.500
0.000
-20
-10
0
10
20
30
40
50
60
Temperature degrees (C)
Si diode
Linear (Si diode)
TINA schematic

Remote System Monitor Applications

Diode temperature monitor

Diode temp sensor circuit

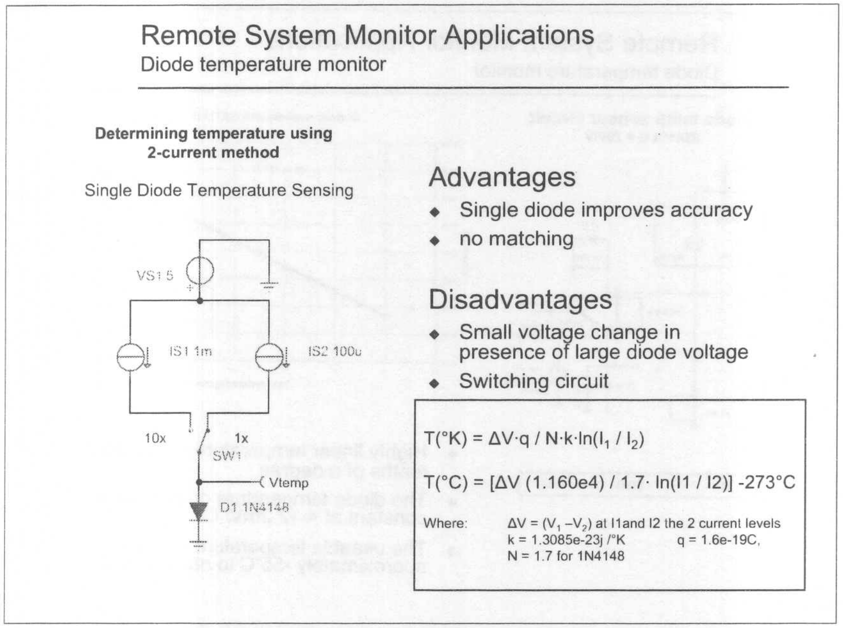
INA114 G = 20V/V

Vtherm= [(VF1 - VF2) (1 + 50kΩ / RG)] + VREF

- Highly linear temperature response – tenths of a degree
- The diode temperature coefficient is constant at ≈ -2.2mV/°C
- The useable temperature range is approximately -55°C to about 175°C

A silicon diode can also be used as a temperature sensor. The forward biased PN junction has voltage temperature coefficient of approximately -2.2mV/°C and is highly consistent over the useable temperature range, -55°C to +175°C. It offers very good linearity performance.

This is basically the same full-bridge circuit that was applied to a thermistor. Note that this circuit doesn't directly measure temperature, but rather temperature change. It must be calibrated using a temperature reference.



Remote System Monitor Applications

Diode temperature monitor

Determining temperature using 2-current method

Single Diode Temperature Sensing

VS1 5
IS1 1m
IS2 100u
10x
1x
SW1
Vtemp
D1 1N4148

Advantages

- Single diode improves accuracy
- no matching

Disadvantages

- Small voltage change in presence of large diode voltage
- Switching circuit

$$T(°K) = \Delta V \cdot q \,/\, N \cdot k \cdot \ln(I_1 / I_2)$$

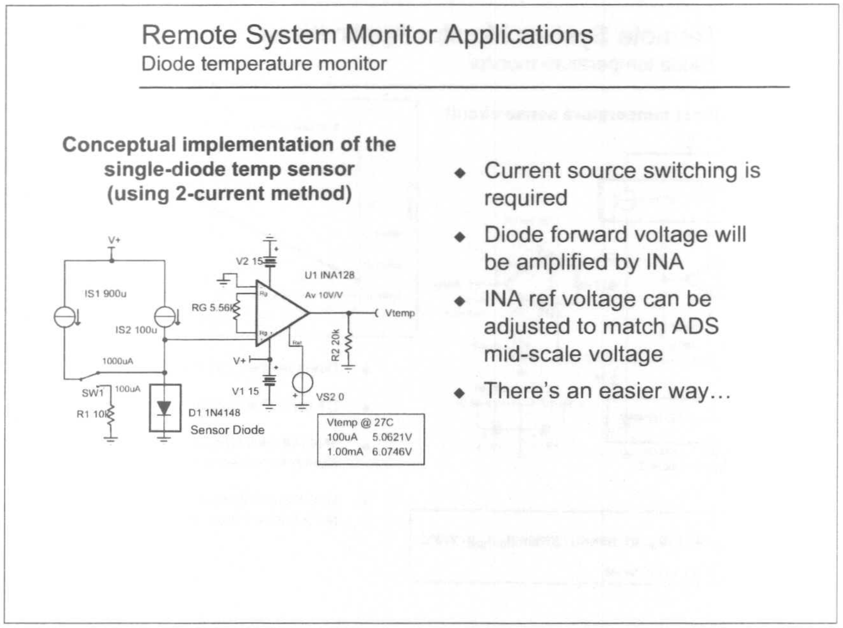
$$T(°C) = [\Delta V\ (1.160e4) \,/\, 1.7 \cdot \ln(I1 / I2)]\ -273°C$$

Where:
$\Delta V = (V_1 - V_2)$ at I1and I2 the 2 current levels
k = 1.3085e-23j /°K   q = 1.6e-19C,
N = 1.7 for 1N4148

A single diode can be used to indicate the actual temperature. The simple circuit presented here shows how this is accomplished.

The diode is subjected to a temperature environment and then the diode forward voltage is measured at 2 significantly different current levels. The current ratio is usually on the order of 10:1 to 100:1. The main consideration with high ratios is the current may be quite high in the high current measurement. This high current can cause the junction to self heat which introduces an error.

The resulting forward voltages are entered into the equation from which temperature is derived.



Remote System Monitor Applications

Diode temperature monitor

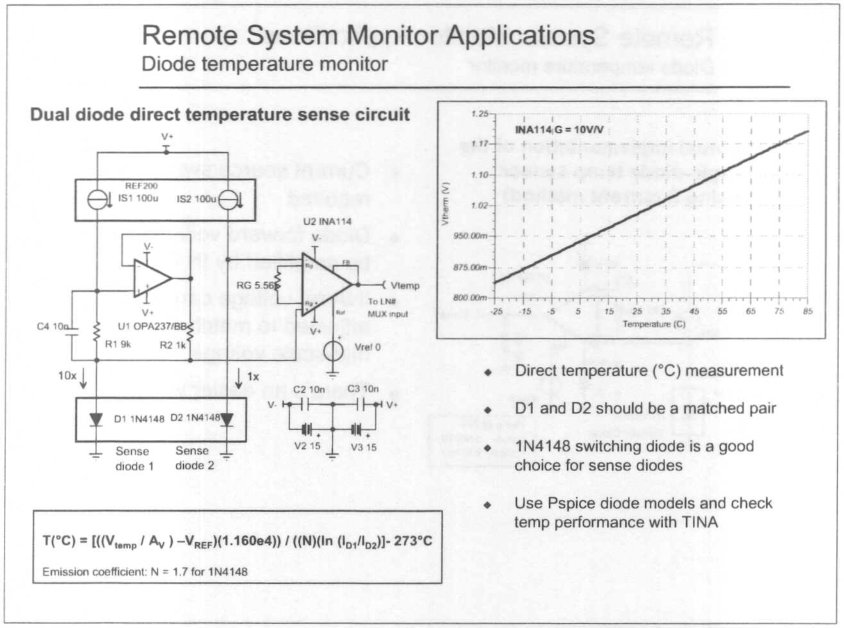
Conceptual implementation of the single-diode temp sensor (using 2-current method)

Vtemp @ 27C	
100uA	5.0621V
1.00mA	6.0746V

- Current source switching is required
- Diode forward voltage will be amplified by INA
- INA ref voltage can be adjusted to match ADS mid-scale voltage
- There's an easier way…

This circuit shows the concept of switching the 2 current levels. An INA114 instrumentation amplifier follows the diode and provides a gain of 10V/V.

The diode's forward voltage is large compared to the voltage change that occurs at the 2 current levels. The instrument amp will gain this relatively large voltage up as well and this results in a large DC value at the output. A reference voltage can be applied to the ref pin to provide an offset so that the instrumentation amplifier's output can be set to the Analog-to-Digital Converter (ADS) mid-scale voltage.



Remote System Monitor Applications

Diode temperature monitor

Dual diode direct temperature sense circuit

- Direct temperature (°C) measurement
- D1 and D2 should be a matched pair
- 1N4148 switching diode is a good choice for sense diodes
- Use Pspice diode models and check temp performance with TINA

$$T(^\circ C) = [((V_{temp} / A_V) - V_{REF})(1.160e4)) / ((N)(\ln (I_{D1}/I_{D2})] - 273^\circ C$$

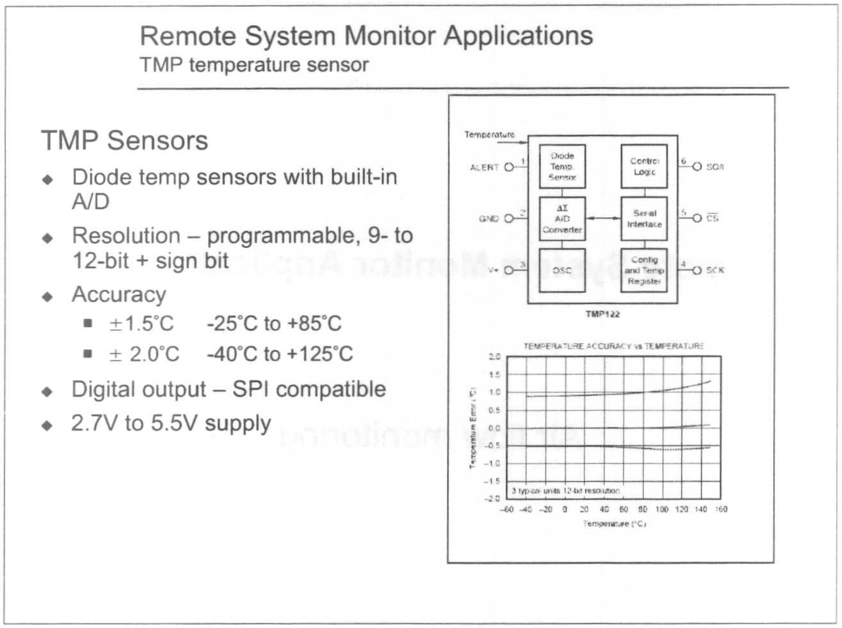
Emission coefficient: N = 1.7 for 1N4148

The requirement to switch currents can be eliminated by using 2 temperature sensing diodes, one operating at 1x current and the other at a higher current such as 10x. The diodes should be matched for the application.

One-half of a REF200 establishes the 1x (100µA) current for one diode. The other half of the REF200 supplies a reference current for a voltage-to-current converter which in turn supplies the 10x (1000µA) current to the other diode.

The INA114 inputs connect to the anode of the two diodes in differential fashion. The INA responds to the difference voltage at these 2 nodes and amplifies it. A reference voltage can be applied to the ref pin to match the INA output with the ADS mid-scale voltage. Also, the INA gain can be adjusted for the appropriate input range.

The temperature can be calculated using the equation.



Remote System Monitor Applications

TMP temperature sensor

TMP Sensors

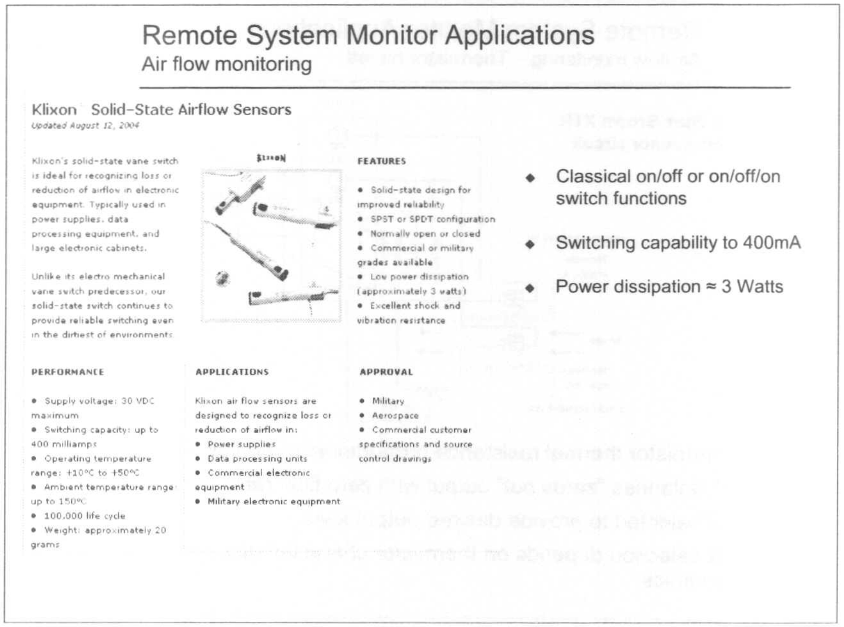
- Diode temp sensors with built-in A/D
- Resolution – programmable, 9- to 12-bit + sign bit
- Accuracy
  - ±1.5°C -25°C to +85°C
  - ± 2.0°C -40°C to +125°C
- Digital output – SPI compatible
- 2.7V to 5.5V supply

Temperature
ALERT
1
GND
2
V+
3
Diode Temp. Sensor
ΔΣ A/D Converter
OSC
Control Logic
Serial Interface
Config and Temp Register
6
SO/I
5
CS
4
SCK
TMP122
TEMPERATURE ACCURACY vs TEMPERATURE
Temperature Error (°C)
2.0
1.5
1.0
0.5
0.0
-0.5
-1.0
-1.5
-2.0
3 typical units 12-bit resolution
-60 -40 -20 0 20 40 60 80 100 120 140 160
Temperature (°C)

The TMP family of integrated temperature sensors offers an easily applied, highly accurate temperature measurement solution. A/D conversion is accomplished by the integrated ΔΣ converter. Therefore, in a remote monitor application, this would be a stand alone function not requiring the ADS7870 A/D function. The SPI compatible, serial output could be directly communicated back to the monitoring station.

Remote System Monitor Applications

Air flow monitoring



Remote System Monitor Applications

Air flow monitoring

Klixon Solid-State Airflow Sensors

Updated August 12, 2004

Klixon's solid-state vane switch is ideal for recognizing loss or reduction of airflow in electronic equipment. Typically used in power supplies, data processing equipment, and large electronic cabinets.

Unlike its electro mechanical vane switch predecessor, our solid-state switch continues to provide reliable switching even in the dirtiest of environments.

FEATURES

- Solid-state design for improved reliability
- SPST or SPDT configuration
- Normally open or closed
- Commercial or military grades available
- Low power dissipation (approximately 3 watts)
- Excellent shock and vibration resistance

PERFORMANCE

- Supply voltage: 30 VDC maximum
- Switching capacity: up to 400 milliamps
- Operating temperature range: +10°C to +50°C
- Ambient temperature range: up to 150°C
- 100,000 life cycle
- Weight: approximately 20 grams

APPLICATIONS

Klixon air flow sensors are designed to recognize loss or reduction of airflow in:

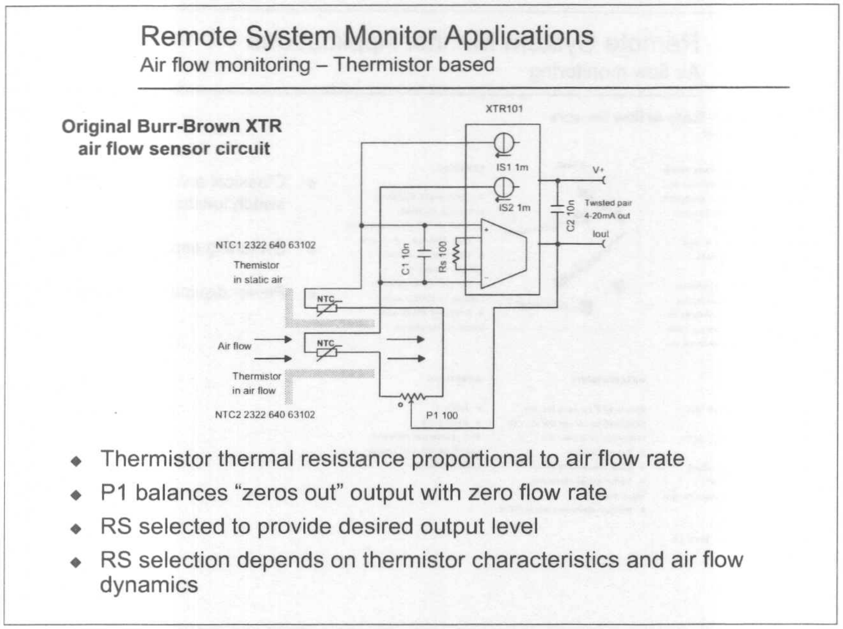
- Power supplies
- Data processing units
- Commercial electronic equipment
- Military electronic equipment

APPROVAL

- Military
- Aerospace
- Commercial customer specifications and source control drawings

- Classical on/off or on/off/on switch functions
- Switching capability to 400mA
- Power dissipation ≈ 3 Watts

This is a classical solid-state, vane type air flow sensor. It provides an on/off or on/off/on switch function.



Remote System Monitor Applications

Air flow monitoring – Thermistor based

Original Burr-Brown XTR air flow sensor circuit

- Thermistor thermal resistance proportional to air flow rate
- P1 balances “zeros out” output with zero flow rate
- RS selected to provide desired output level
- RS selection depends on thermistor characteristics and air flow dynamics

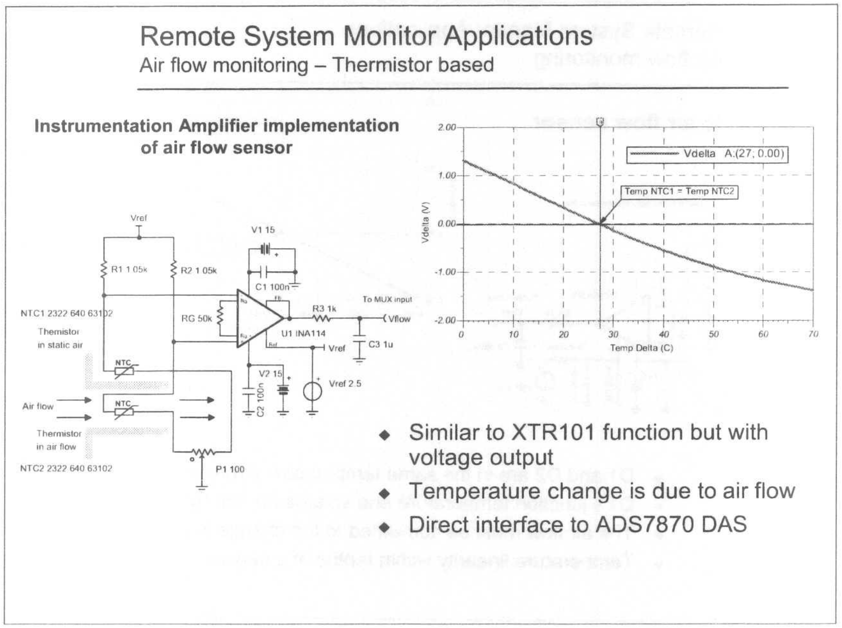
The air flow monitor shown here is an outgrowth of the temperature monitors presented in the previous section. A thermistor serves as the air flow sense element. This circuit uses an XTR101, 4-20mA current loop transmitter implementation originally presented in Burr-Brown applications bulletin AB-032A.

Two thermistors are used in the bridge. One thermistor is located in the still environment while the other is placed in the air stream. The thermistor environments should be at the same temperature so as not to introduce a temperature gradient error. Both thermistors have a small bias current flowing through them and self heat to a small degree. This is an important point to keep in mind.

Air flowing past will add or remove heat to the exposed thermistor, changing its temperature relative to the thermistor in the static air environment. The amount of temperature change will be related to the air flow dynamics.

The change in thermistor temperature leads to a resistance change that imbalances the bridge, which in turn appears as a voltage difference to the instrumentation amplifier inputs.

An XTR can be used in remote system monitoring applications. However, for use with the ADS7870 DAS, the current output (4-20mA) would have to be converted to a voltage.



Remote System Monitor Applications

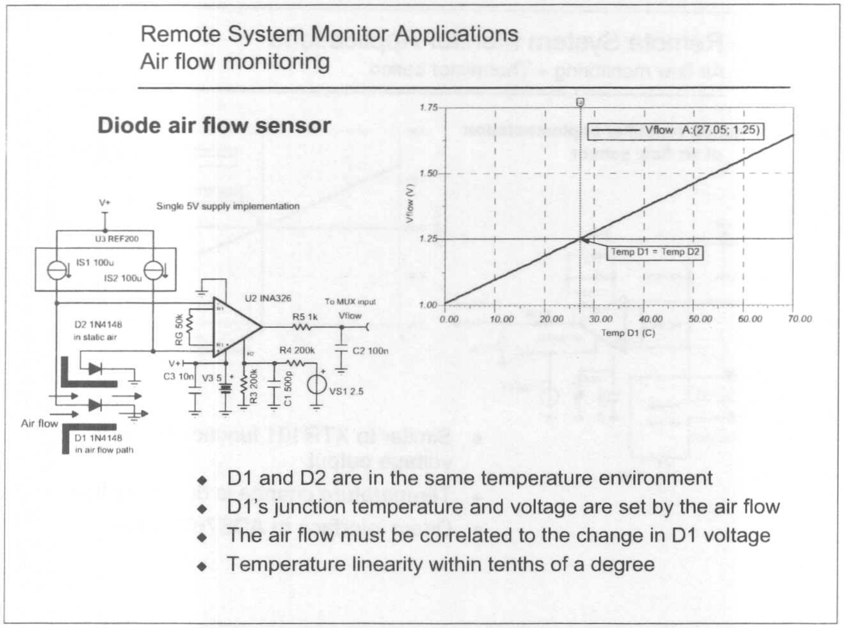
Air flow monitoring – Thermistor based

Instrumentation Amplifier implementation of air flow sensor

- Similar to XTR101 function but with voltage output
- Temperature change is due to air flow
- Direct interface to ADS7870 DAS

This is an implementation of the previous thermistor bridge, but now used in conjunction with an instrumentation amplifier. The instrumentation amplifier provides a voltage output which can be directly interfaced with the ADS7870 DAS.

The plot shows the delta voltage relative to the Vref voltage. The voltage change is directly related to the heating or cooling of the sense response to the air flow. The “delta” is zero when the temperature of the two thermistors is exactly the same. Air flow will cause a temperature delta in accordance with the plot.



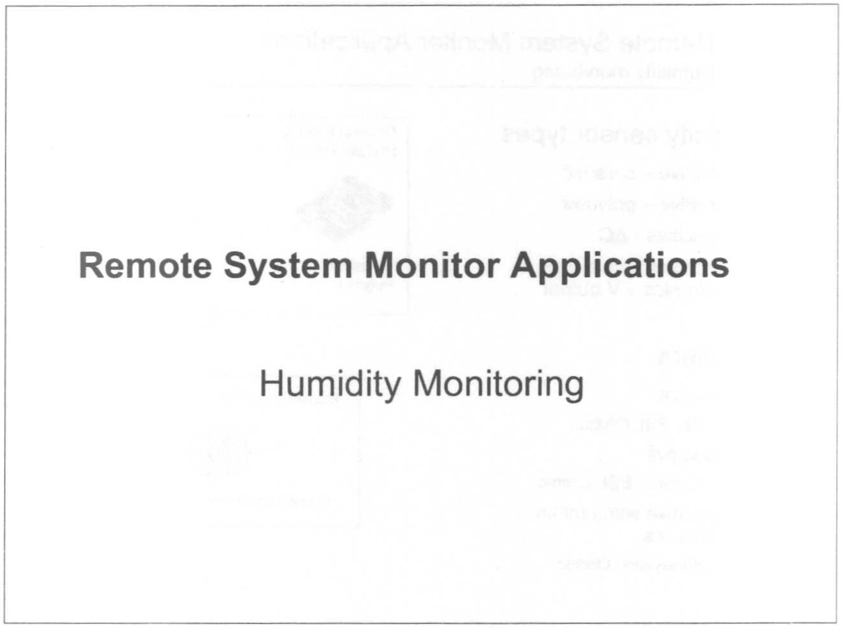
Remote System Monitor Applications
Air flow monitoring

Diode air flow sensor

- D1 and D2 are in the same temperature environment
- D1's junction temperature and voltage are set by the air flow
- The air flow must be correlated to the change in D1 voltage
- Temperature linearity within tenths of a degree

As previously mentioned, the silicon diode exhibits a linear junction voltage change with temperature of approximately -2.2mV/°C. This was shown to be a direct indicator of junction temperature change. In similar fashion to the thermistor, and by virtue of the junction's voltage change in response to heating or cooling, the diode can be used as an air flow sensor.

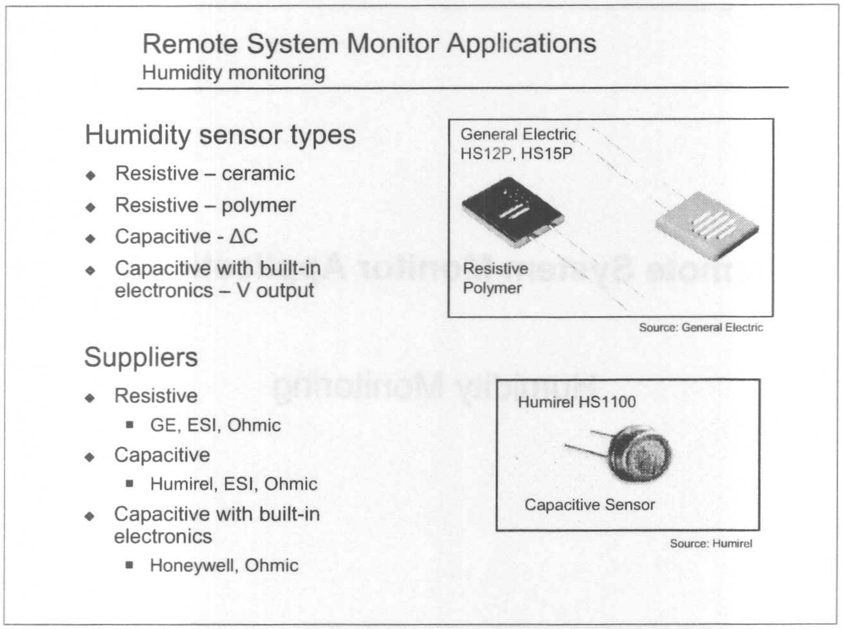
The air flow dynamics dictate how much the voltage will increase or decrease for a given flow rate. Therefore, the relationship between temperature and flow rate would have to be established for a method such as this to be useable.



Remote System Monitor Applications

Humidity Monitoring

Humidity sensors are typically resistive or capacitive sensors. Their characteristics are much different than the previously discussed sensors and require a different approach in converting their output to something that can interface directly with the ADS7870 DAS.



Remote System Monitor Applications

Humidity monitoring

Humidity sensor types

- Resistive – ceramic
- Resistive – polymer
- Capacitive - ΔC
- Capacitive with built-in electronics – V output

General Electric
HS12P, HS15P

Resistive
Polymer

Source: General Electric

Suppliers

- Resistive
  - GE, ESI, Ohmic
- Capacitive
  - Humirel, ESI, Ohmic
- Capacitive with built-in electronics
  - Honeywell, Ohmic

Humirel HS1100

Capacitive Sensor

Source: Humirel

Resistive humidity sensors usually consist of a hygroscopic (absorbs moisture) medium such as conductive salt or polymer deposited over noble metal electrodes on a nonconductive substrate.

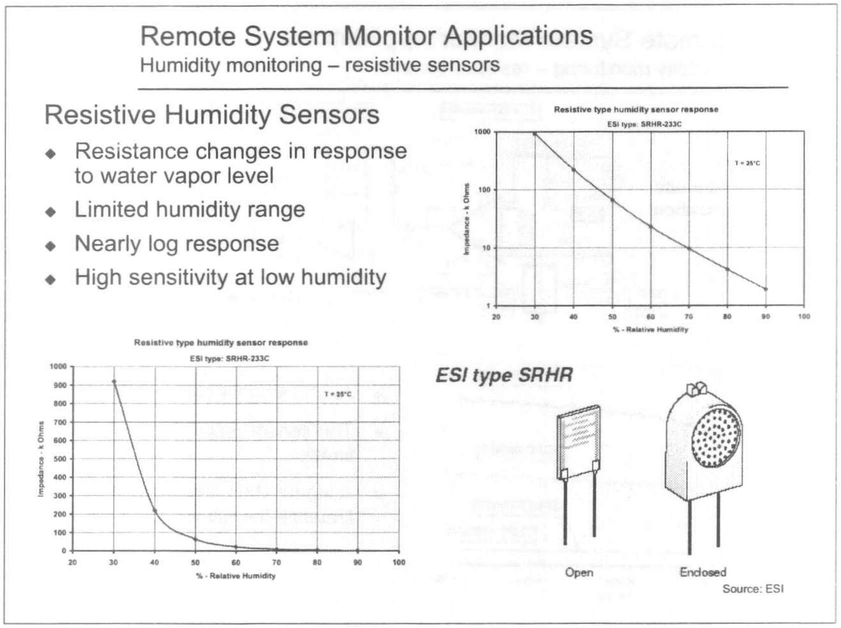
When the sensor is in the presence of water vapor, the vapor is absorbed causing the functional ionic groups to disassociate, resulting in increased conductivity. Response times are slow ranging from 10 to 30s for a 63% step change.

Most resistive sensors use an AC excitation to prevent sensor polarization. The resulting current is rectified and converted to DC where it can then undergo linearization and be amplified as necessary. The AC signal applied to the bridge ranges from 30Hz to 10kHz.

The capacitive sensor is constructed of a thin polymer or metal oxide deposited between two conductive plates on a ceramic, glass or silicon substrate. The sensing surface is then coated with a porous metal coating to protect it from contamination and exposure to condensation. An incremental change in the dielectric constant of the dielectric takes places in the presence of moisture.

A 3rd humidity sensor type is based on a 2-thermistor design. One thermistor is sealed in dry nitrogen and serves as the reference, while the other is exposed to the ambient air. The “dry” thermistor has a greater capacity to sink or release heat than the exposed “wet” thermistor. The thermistors are biased to self heating levels. Since their ability to dissipate heat is different the resistance of each will take on a unique value that unbalances the bridge.

(source sensormag.com/articles)



Remote System Monitor Applications

Humidity monitoring – resistive sensors

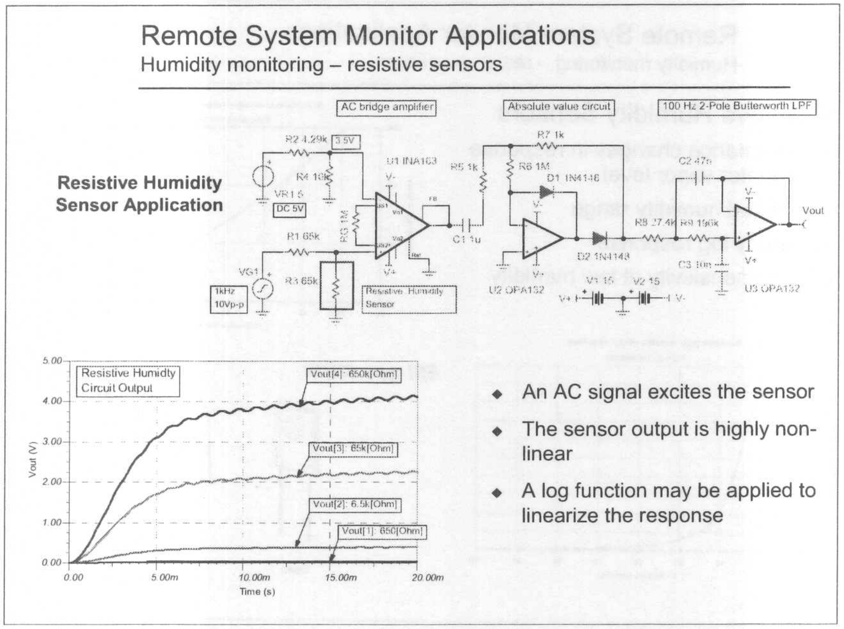
Resistive Humidity Sensors

- Resistance changes in response to water vapor level
- Limited humidity range
- Nearly log response
- High sensitivity at low humidity

Resistive type humidity sensor response
ESI type: SRHR-233C
Impedance - k Ohms
1000
100
10
1
T = 25°C
20
30
40
50
60
70
80
90
100
% - Relative Humidity
Resistive type humidity sensor response
ESI type: SRHR-233C
Impedance - k Ohms
1000
900
800
700
600
500
400
300
200
100
0
T = 25°C
20
30
40
50
60
70
80
90
100
% - Relative Humidity
ESI type SRHR
Open
Enclosed

Source: ESI

The resistive type humidity sensor has a very nonlinear response for a uniform change in relative humidity. When the points are re-plotted on a logarithmic scale the response has a more linear appearance.



Remote System Monitor Applications

Humidity monitoring – resistive sensors

Resistive Humidity Sensor Application

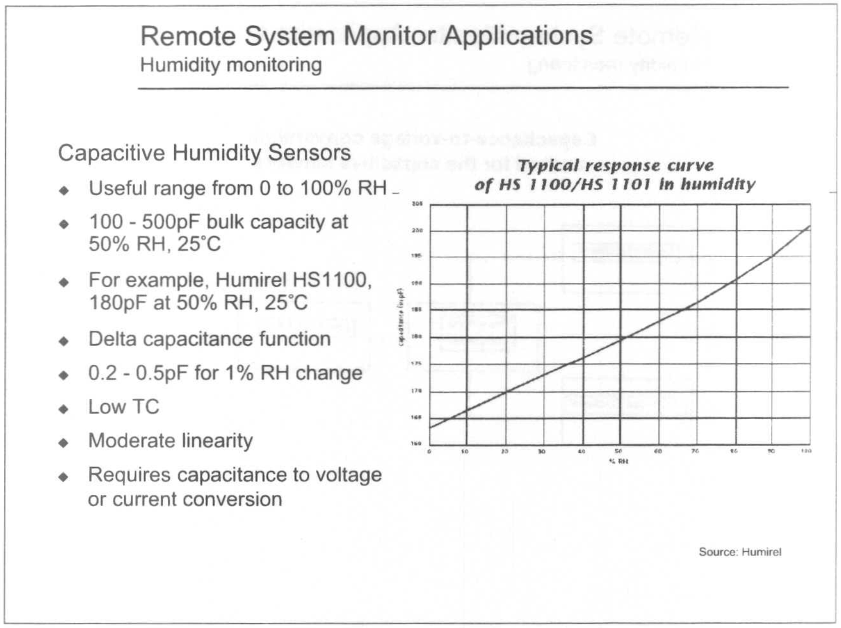
- An AC signal excites the sensor
- The sensor output is highly non-linear
- A log function may be applied to linearize the response

A 1kHz AC sine wave is used to excite a half-bridge which includes the resistive humidity sensor. As can be seen from the graphs the output voltage is proportional to the voltage divider voltage established by the upper resistor and resistive sensor. The resistive humidity sensor is excited with an AC signal, instead of a DC voltage, to prevent sensor polarization.

A DC pedestal voltage is applied to the inverting input by way of a voltage divider. This DC voltage is subtracted from the AC voltage by the instrumentation amplifier and a difference voltage appears at the output. The DC level assures that the difference voltage at the two inputs results in an increasing output voltage as the voltage difference increases.

The output of the AC bridge amplifier is AC coupled, to the absolute value circuit which performs a full-wave rectifier function. The rectified voltage is then applied to the input of a 2-pole, low-pass filter. This removes most of the 1kHz ripple.

Since the humidity sensor produces a highly non-linear resistance change with a linear change in humidity level, the output voltage is non-linear as well. A log amplifier could be employed to help improve the output voltage versus humidity response.



Remote System Monitor Applications

Humidity monitoring

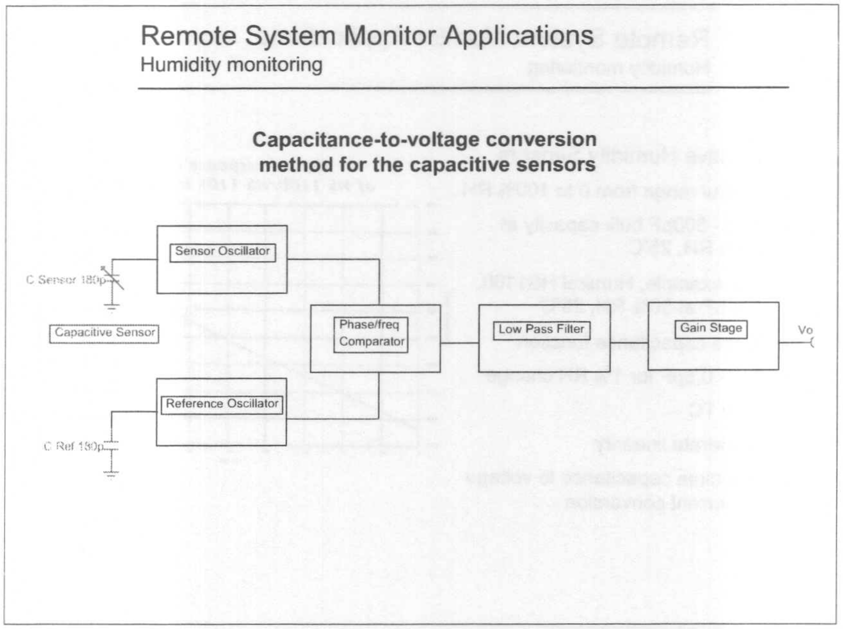
Capacitive Humidity Sensors

- Useful range from 0 to 100% RH
- 100 - 500pF bulk capacity at 50% RH, 25°C
- For example, Humirel HS1100, 180pF at 50% RH, 25°C
- Delta capacitance function
- 0.2 - 0.5pF for 1% RH change
- Low TC
- Moderate linearity
- Requires capacitance to voltage or current conversion

Typical response curve of HS 1100/HS 1101 in humidity

Source: Humirel

The capacitive humidity sensor has a more linear response than the resistive sensor. It is also usable over the entire range of 0 to 100% relative humidity, where the resistive element is limited to about 20 to 90% relative humidity.



Remote System Monitor Applications

Humidity monitoring

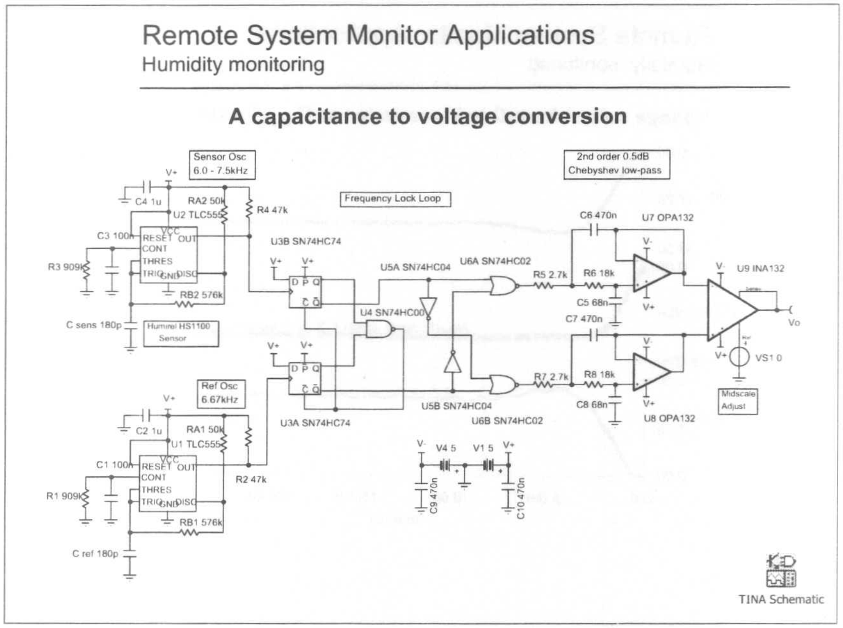
Capacitance-to-voltage conversion method for the capacitive sensors

Capacitive sensors produce a capacitance change, a “delta-C,” in response to a change in the monitored attribute. This capacitance change is then often converted to a different electrical property that can be more directly measured. Often this is a voltage which necessitates a capacitance-to-voltage (C-V) conversion.

One technique employed for the C-V conversion involves placing the capacitive sensor in the frequency determining circuit of an oscillator. Then, any capacitance change will alter the oscillator frequency.

In the accompanying diagram a reference oscillator with a fixed frequency and a sensor oscillator with a variable frequency are shown. The output signals from the two oscillators are then compared by a phase or frequency comparator, sometimes referred to as a phase discriminator.

The phase/frequency discriminator will produces a DC level or pulse-width modulated (PWM) pulse train that is a function of the phase or frequency difference of the two oscillator signals - depending on the design. The output is then applied to a low-pass filter, to filter or integrate the discriminator output. It may then be amplified by a gain stage as needed.



Remote System Monitor Applications

Humidity monitoring

A capacitance to voltage conversion

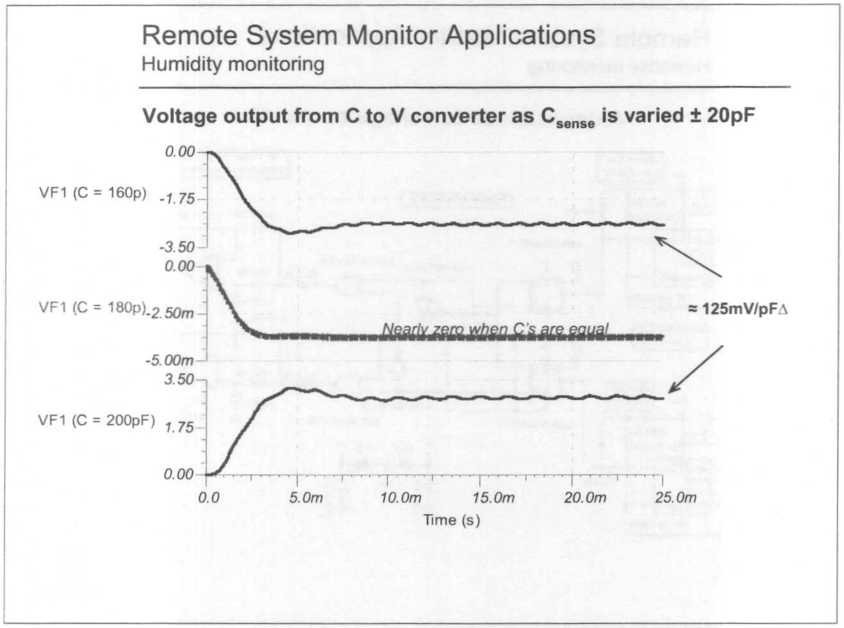
TINA Schematic

Here is an implementation of the C-V conversion just described. It uses two TLC555, bi-stable multivibrators as the oscillators and a frequency lock loop (FLL), as a frequency difference detector.

The reference oscillator has a fixed frequency of 6.7kHz, while the sensor oscillator has a frequency that changes from 6.0 to 7.5kHz with a ±20pF sensor capacitance change.

If the sensor oscillator frequency moves relative to the reference frequency one or the other NOR gate in the FLL begins to output a series of pulses whose width increases with larger frequency differences. This is essentially a pulse width modulated (PWM), pulse train. When the oscillators are operating at the same frequency the FLL does not produce an output.

The pulse train then passes through the second-order low pass filter, which serves as an integrator. The integrator function produces a DC level in proportion to the pulse width. Then, the differential amplifier amplifies the differential DC level while rejecting any common responses.



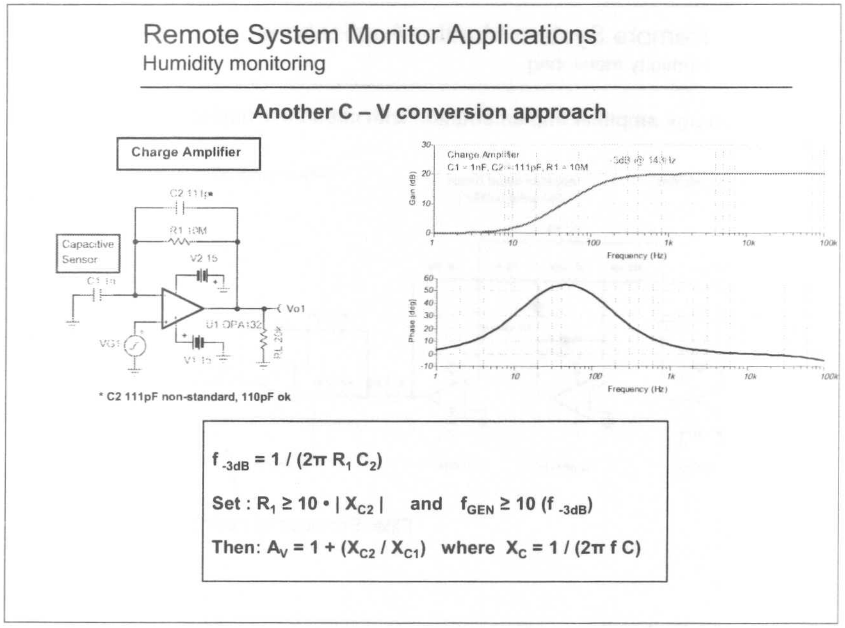
Remote System Monitor Applications

Humidity monitoring

Voltage output from C to V converter as C_{sense} is varied ± 20pF

This is the output response for the C-V circuit with a ±20pF capacitance change. The center plot shows that the output voltage is nearly zero when the 2 oscillators are on the same frequency.

However, if the sensor oscillator is at either one of the frequency extremes, then the output voltage moves off zero to approximately ± 2.5V. The circuit exhibits excellent sensitivity, about 125mV/pF.



Remote System Monitor Applications

Humidity monitoring

Another C – V conversion approach

Charge Amplifier
C2 111p*
R1 10M
Capacitive Sensor
V2 15
C1 1n
Vo1
U1 OPA132
VG1
V1 15
RL 20k
30
20
10
0
Gain (dB)
Charge Amplifier
C1 = 1nF, C2 = 111pF, R1 = 10M
-3dB @ 143Hz
1
10
100
1k
10k
100k
Frequency (Hz)
60
50
40
30
20
10
0
-10
Phase (deg)
Frequency (Hz)

* C2 111pF non-standard, 110pF ok

$f_{-3dB} = 1 / (2\pi R_1 C_2)$

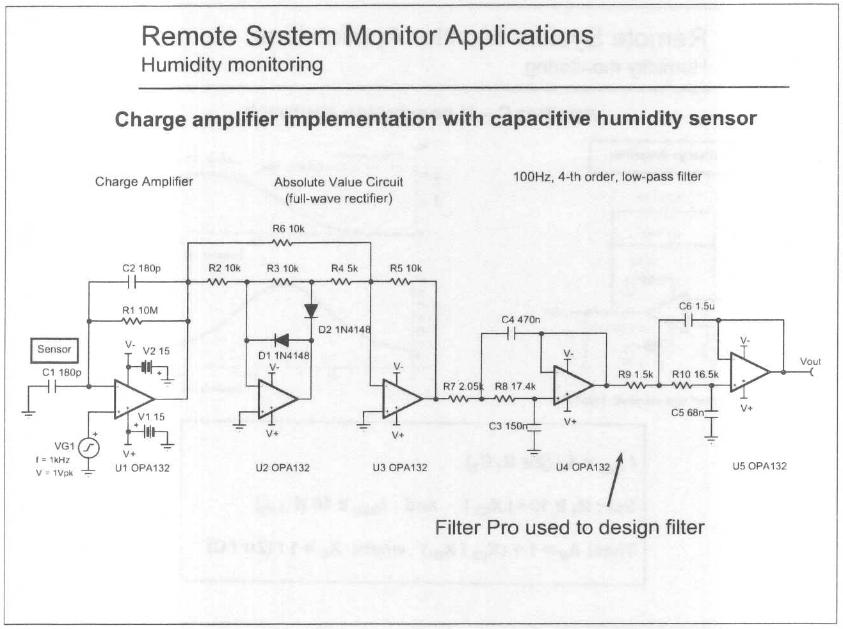
Set : $R_1 \geq 10 \cdot | X_{C2} |$ and $f_{GEN} \geq 10 (f_{-3dB})$

Then: $A_V = 1 + (X_{C2} / X_{C1})$ where $X_C = 1 / (2\pi f C)$

Another approach to C-V conversion is the Charge Amplifier. Here, the op-amp, closed-loop voltage gain (Av) is a ratio function of the capacitive reactances ($-jX_C$) in the feedback and inverting input circuits.

If either reactance changes the Av of the amplifier will change as well. A carrier signal applied to the non-inverting input will be amplified, but by different gain levels depending on the reactance ratio.

This varying amplitude AC signal can then be rectified to extract a DC voltage level.



Remote System Monitor Applications

Humidity monitoring

Charge amplifier implementation with capacitive humidity sensor

Charge Amplifier

Absolute Value Circuit
(full-wave rectifier)

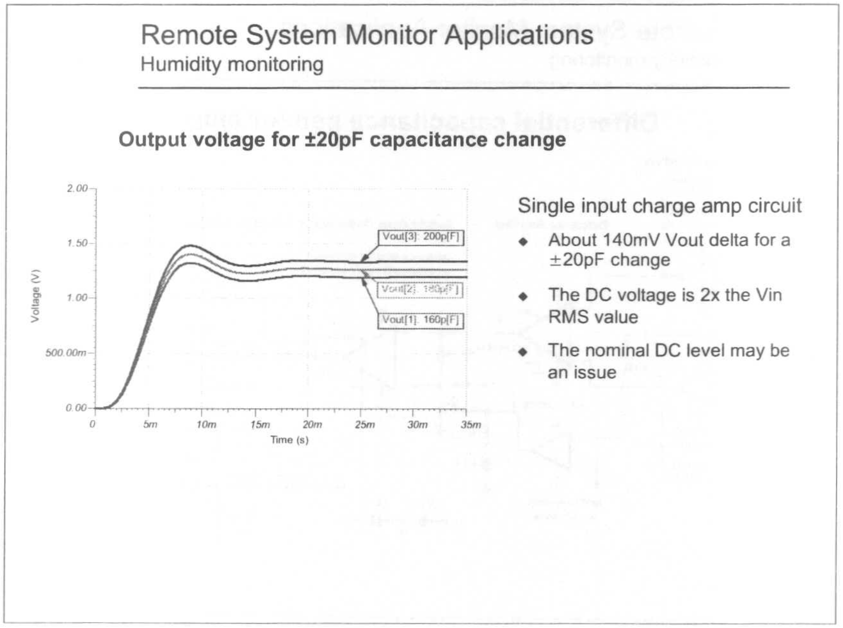
100Hz, 4-th order, low-pass filter

Filter Pro used to design filter

This is a complete charge amplifier circuit useful for detection of a capacitive sensor's capacitance change.

The charge amplifier is followed by an absolute value circuit which serves as a full-wave rectifier. This converts the varying AC signal from the charge amplifier to a DC level. However, the DC level is unfiltered at this point and contains the carrier frequency ripple.

The rectified DC is then passed through a 2-stage, 4-pole low pass filter to remove the ripple. This results in a DC voltage that is a function of the charge amplifier gain. TI's Filter Pro tool was used to design this 4th order filter.



Remote System Monitor Applications

Humidity monitoring

Output voltage for ±20pF capacitance change

Single input charge amp circuit

- About 140mV Vout delta for a ±20pF change
- The DC voltage is 2x the Vin RMS value
- The nominal DC level may be an issue

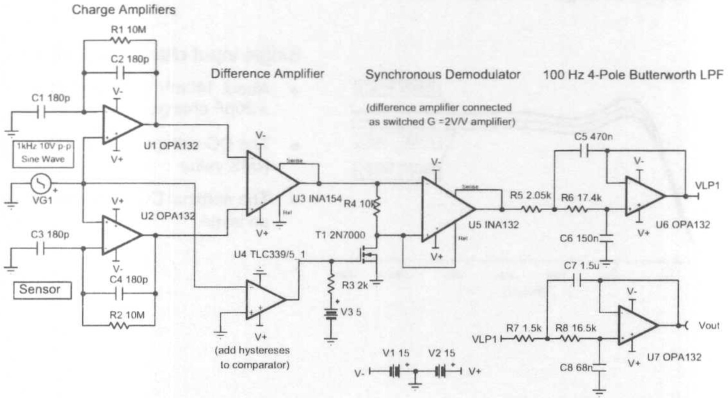
The previous charge amplifier circuit produces a stable DC level for a fixed charge amp gain. Here the input sensor capacitance is changed to 3 different values; 160, 180 and 200pF, resulting in three different charge amplifier gains and corresponding DC levels.

The circuit is not overly sensitive and a large DC level is present. Adding gain would help exaggerate the DC change that results from the capacitance change, but the large DC level would be gained up as well. This would have to be dealt with in a subsequent stage.

Remote System Monitor Applications

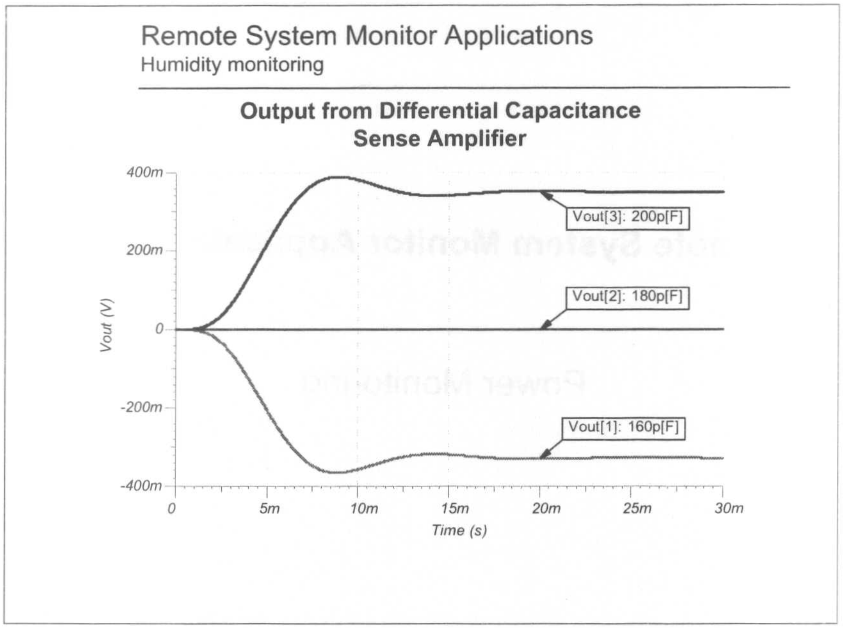
Humidity monitoring

Differential capacitance sensor amplifier



A unique differential implementation of the charge amplifier is shown here. The output from the sensor charge amplifier is compared to that of a reference charge amplifier. Differential signals applied to the differential amplifier (U3) inputs are amplified, while common mode signals are rejected. This removes the large common-mode DC voltage had with the previous, single-ended charge amplifier circuit.

The resulting AC output is converted to a rectified DC level by a synchronous demodulator consisting of U4, T1 and U5. As with the earlier circuit, carrier ripple is present and must be filtered. The 100Hz, 4-pole, Butterworth filter accomplishes this task.



Remote System Monitor Applications

Humidity monitoring

Output from Differential Capacitance Sense Amplifier

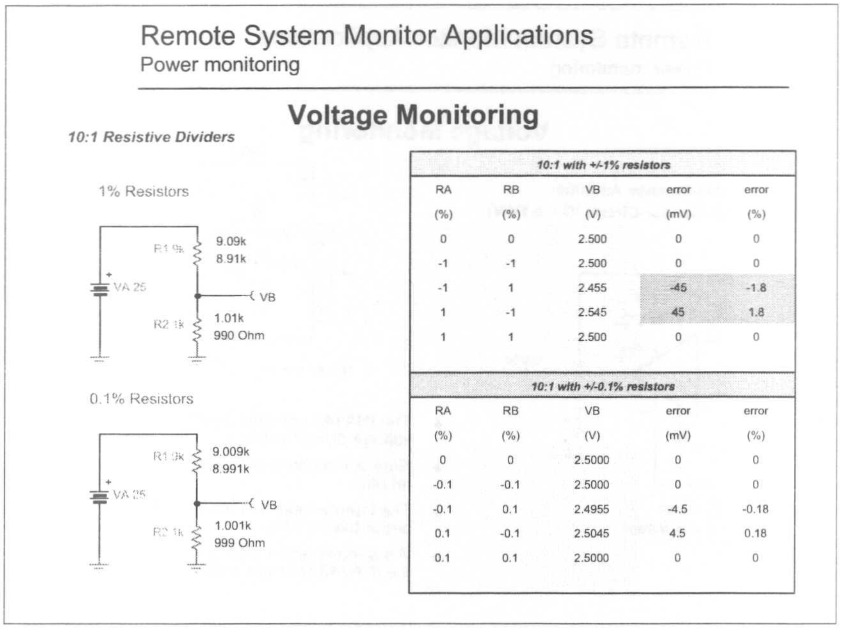
The output from the differential capacitance sensor amplifier is shown in the accompanying graph. The plots show that after approximately 20ms the DC level stabilizes to a final value. When the charge amplifier capacitances are equal the circuit is balanced and the output is zero.

When the sensor capacitance is at its minimum (nominal -20pF) the output voltage is approximately -350mV. The output voltage is opposite and equal at +350mV when the sensor capacitance is at its maximum (nominal +20pF).

Remote System Monitor Applications

Power Monitoring

Voltage and current levels can be easily and precisely monitored using analog techniques.



Remote System Monitor Applications

Power monitoring

Voltage Monitoring

10:1 Resistive Dividers

1% Resistors

0.1% Resistors

10:1 with +/-1% resistors

RA (%)	RB (%)	VB (V)	error (mV)	error (%)
0	0	2.500	0	0
-1	-1	2.500	0	0
-1	1	2.455	-45	-1.8
1	-1	2.545	45	1.8
1	1	2.500	0	0

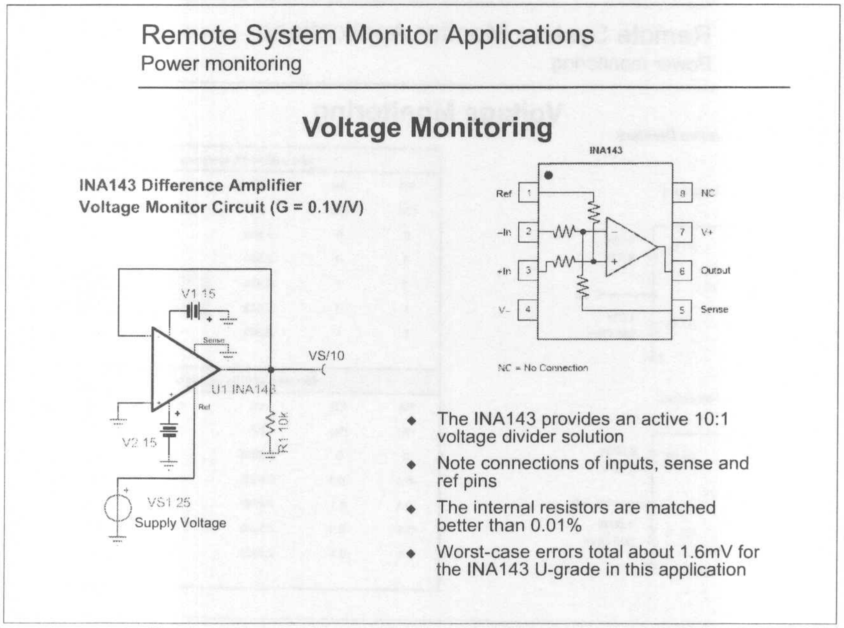
10:1 with +/-0.1% resistors

RA (%)	RB (%)	VB (V)	error (mV)	error (%)
0	0	2.5000	0	0
-0.1	-0.1	2.5000	0	0
-0.1	0.1	2.4955	-4.5	-0.18
0.1	-0.1	2.5045	4.5	0.18
0.1	0.1	2.5000	0	0

The DAS is capable of handling input voltages of 0 to 5V when powered with a 5V supply and a 2.5 reference voltage is employed. Often, the voltage exceeds this level and simple voltage divider can be inserted before the DAS to reduce the voltage to a level it can safely handle.

This slide provides an example where the supply voltage is 25V and it is desired to divide that by a factor of 10. Two dividers are shown in the example and the resistor tolerances at ±1% and ±0.1%, respectively.

Depending on the direction of each resistor's tolerance it is seen that resistors with a 1% tolerance can result in a divider error as high as ±1.8%. Similarly, the 0.1% tolerance resistors may result in an error as high as ±0.18%. Either may be acceptable. It just depends on the system requirements.



Remote System Monitor Applications

Power monitoring

Voltage Monitoring

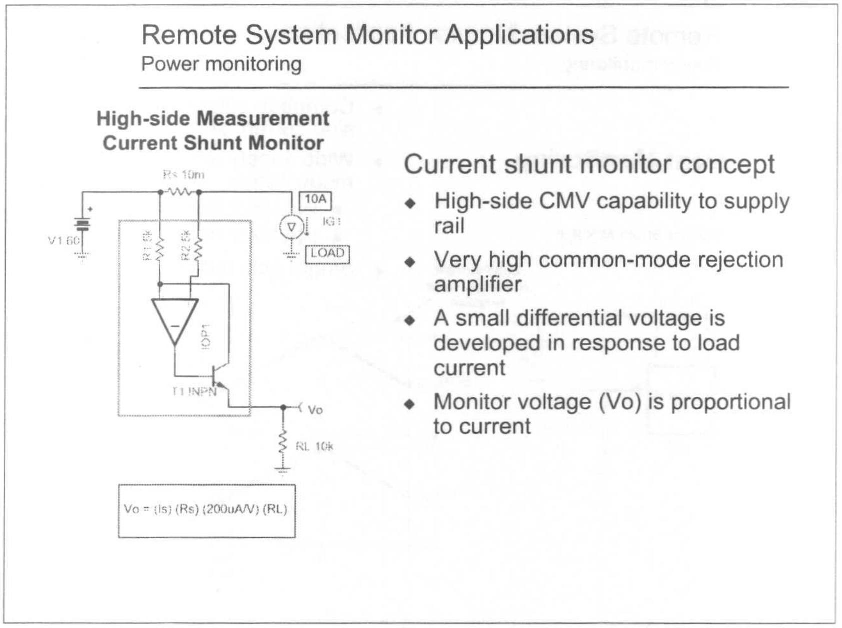
INA143 Difference Amplifier Voltage Monitor Circuit (G = 0.1V/V)

- The INA143 provides an active 10:1 voltage divider solution
- Note connections of inputs, sense and ref pins
- The internal resistors are matched better than 0.01%
- Worst-case errors total about 1.6mV for the INA143 U-grade in this application

The INA143 may be configured in a manner such that it will provide a precise, 10:1, voltage-divider function which is useful for monitoring voltage. This is accomplished by reversing the feedback and input resistor on the inverting input and the divider resistors on the non-inverting input which results in a gain of 0.1V/V.

One might wonder why one would resort to this approach. Mostly, it comes down to the accuracy required and the cost. One-percent resistors cost about 10 cents or less, in quantity, at this time. One-tenth percent resistors cost about a half dollar to over a dollar, and 0.01% resistors cost $5 to $10 a piece. And 2 resistors are required for the voltage divider.

Considering the INA143 internal resistors are matched to better than 0.01%, at a cost around $1.50, it offers an accurate, buffered, cost effective solution.



Remote System Monitor Applications

Power monitoring

High-side Measurement
Current Shunt Monitor

Rs 10m
10A
IG1
LOAD
V1 60
R1 5k
R2 5k
IOP1
T1 NPN
Vo
RL 10k

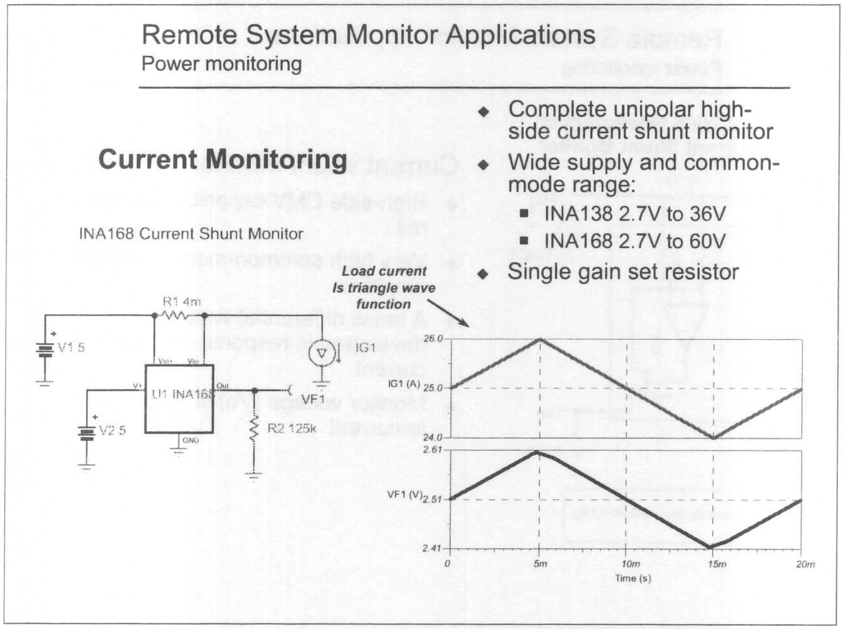
Vo = (Is) (Rs) (200uA/V) (RL)

Current shunt monitor concept

- High-side CMV capability to supply rail
- Very high common-mode rejection amplifier
- A small differential voltage is developed in response to load current
- Monitor voltage (Vo) is proportional to current

The high-side measurement, current shunt monitor provides a convenient means to monitor a load current. In a sense, the circuit performs a function similar to a classical ammeter. In a classical ammeter circuit a sensitive voltmeter, with a full-scale range of 50 to 100mV, is shunted by a very low value current shunt resistor. The meter voltage is proportional to the current flowing through the shunt resistor.

A high CMV tolerant operational amplifier replaces the meter with the current shunt monitor IC. Current flowing though the shunt resistor produces a differential voltage that is amplified by the op-amp and scaled as needed to indicate the current magnitude.



Remote System Monitor Applications

Power monitoring

Current Monitoring

- ◆ Complete unipolar high-side current shunt monitor
- ◆ Wide supply and common-mode range:
  - INA138 2.7V to 36V
  - INA168 2.7V to 60V
- ◆ Single gain set resistor

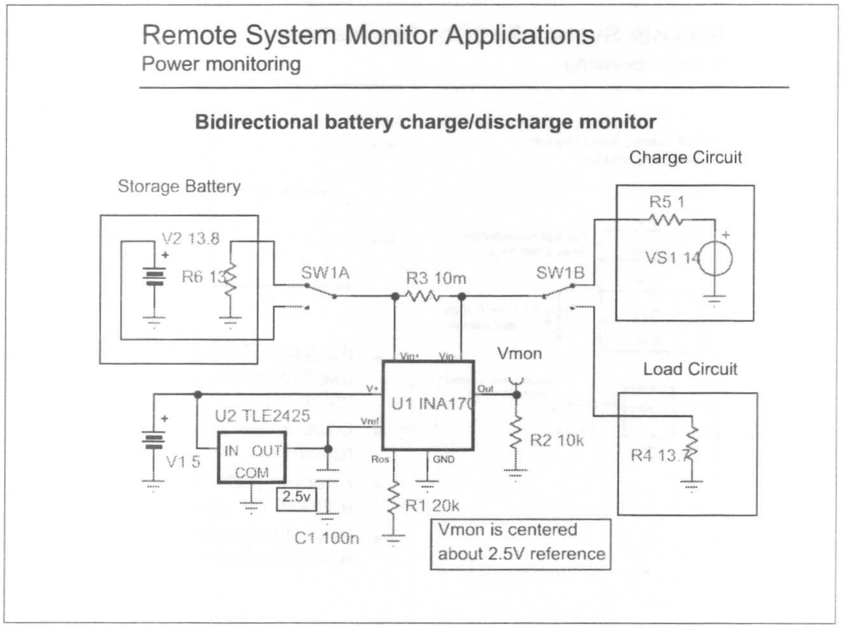
INA168 Current Shunt Monitor

Load current Is triangle wave function

R1 4m
V1 5
+
IG1
Vin+
Vin-
V+
U1 INA168
Out
VF1
V2 5
+
R2 125k
GND
26.0
IG1 (A) 25.0
24.0
2.61
VF1 (V) 2.51
2.41
0
5m
10m
15m
20m
Time (s)

The INA138 and INA168 are examples of high-side measurement, current shunt monitor ICs. This TINA example shows that they can be employed not only in DC but AC applications as well.

The common-mode voltage (CMV) input range is independent of the supply voltage. This high voltage capability of the shunt monitor allows input voltages well above the supply voltage to be monitored.



Remote System Monitor Applications

Power monitoring

Bidirectional battery charge/discharge monitor

Vmon is centered about 2.5V reference

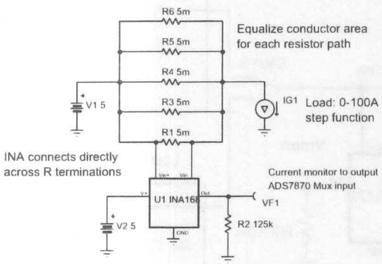
The INA170 is a high-side, bidirectional current shunt monitor IC that is useful for applications such as battery charge and discharge rate monitoring and power management in portable devices.

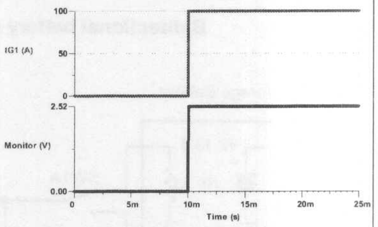
The bidirectional current measurement capability is made possible by output offsetting. This offset is established with an external resistor, Ros, and an external reference voltage. Doing so allows the INA170 to be powered by a single power supply.

Remote System Monitor Applications

Power monitoring

INA168 Current Shunt Monitor
100Amp Application

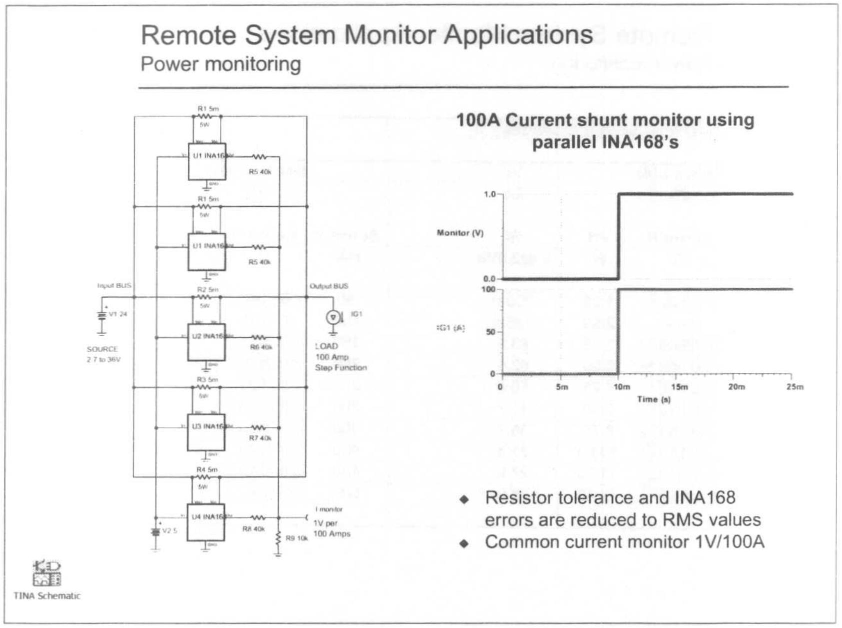

R6 5m
R5 5m
R4 5m
R3 5m
R1 5m
Equalize conductor area for each resistor path
V1 5
IG1
Load: 0-100A step function
INA connects directly across R terminations
Vin+
Vin-
V+
U1 INA168
Out
GND
Current monitor to output ADS7870 Mux input
VF1
R2 125k
V2 5



- Paralleling resistors can be used to achieve very low values R's
- Uses available 5mΩ resistors
- Keeping path Ω's equal and low may be difficult
- Heat dissipation must be accommodated

When monitoring very large current levels, hundreds-of-amperes or more, it may be difficult to find suitable, high power shunt resistors. Shunt resistors with values below 5mΩ are less available than higher value resistors. One option is to split the shunt resistor up among several resistors connected in parallel. In this example the 1mΩ resistance is satisfied by 5 equal value, paralleled resistors, each with 1/5 the power rating requirements of a single resistor.

There is some risk that unequal resistor contact resistance could affect the current through each resistor path.



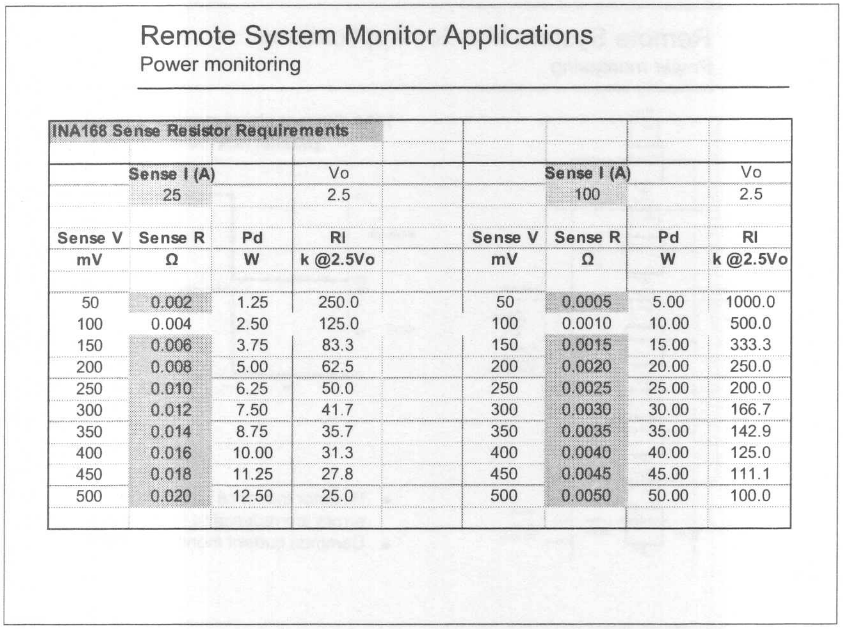
Remote System Monitor Applications

Power monitoring

100A Current shunt monitor using parallel INA168's

- Resistor tolerance and INA168 errors are reduced to RMS values
- Common current monitor 1V/100A

A separate INA168 current shunt monitor has been assigned to each of the 5 shunt resistor current paths. The circuit has the advantage over the previous in that errors such as resistor tolerance and those associated with the IN168's, gain, offset, noise, drift, etc, will now be the statistical average, or root mean squared (RMS) value of all the components. This, in general, will be less than most of the like component population.

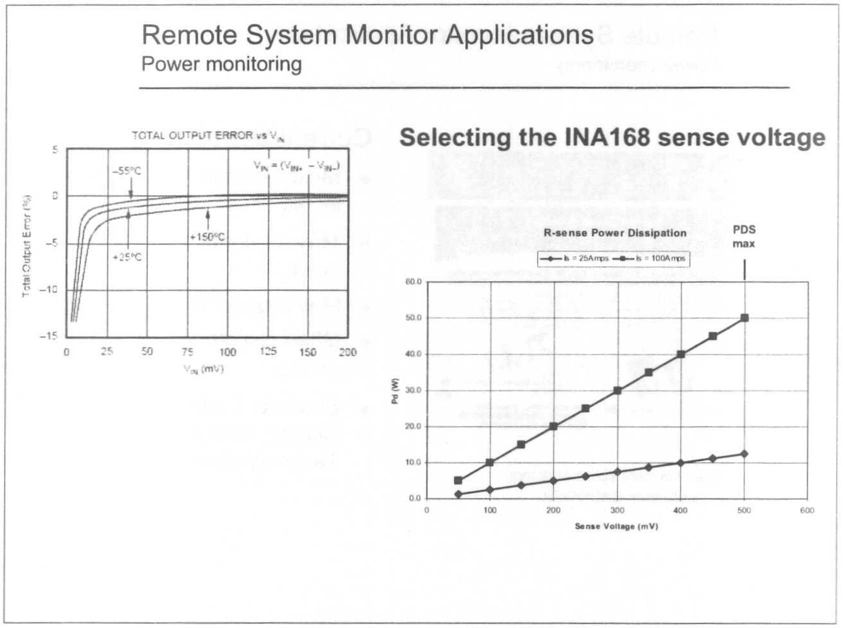


Remote System Monitor Applications

Power monitoring

INA168 Sense Resistor Requirements								
	Sense I (A)		Vo			**Sense I (A)**		Vo
	25		2.5			100		2.5
Sense V mV	**Sense R Ω**	**Pd W**	**RI k @2.5Vo**		**Sense V mV**	**Sense R Ω**	**Pd W**	**RI k @2.5Vo**
50	0.002	1.25	250.0		50	0.0005	5.00	1000.0
100	0.004	2.50	125.0		100	0.0010	10.00	500.0
150	0.006	3.75	83.3		150	0.0015	15.00	333.3
200	0.008	5.00	62.5		200	0.0020	20.00	250.0
250	0.010	6.25	50.0		250	0.0025	25.00	200.0
300	0.012	7.50	41.7		300	0.0030	30.00	166.7
350	0.014	8.75	35.7		350	0.0035	35.00	142.9
400	0.016	10.00	31.3		400	0.0040	40.00	125.0
450	0.018	11.25	27.8		450	0.0045	45.00	111.1
500	0.020	12.50	25.0		500	0.0050	50.00	100.0

This table provides the details for the sense resistors required for specific sense voltages, at the 2 different current levels; 25A and 100A. Keeping the INA168 sense voltage to 50 or 100mV results in lower power dissipation, but also at very high current levels results in minute resistor values which may be difficult to realize.



Remote System Monitor Applications

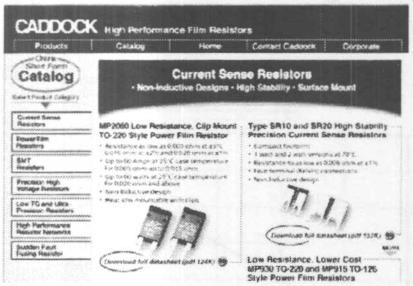
Power monitoring

Selecting the INA168 sense voltage

There is a tradeoff between minimizing the INA168 total output error and the sense resistor power dissipation. Increasing the sense voltage minimizes the errors (to a point), but power dissipation in the sense resistor increases. Even though the lowest output errors are attained by using a larger sense voltage, there is little reason for using a sense voltage above 200mV - even though the product has a 500mV maximum specification.

Remote System Monitor Applications

Power monitoring



CADDOCK High Performance Film Resistors
Products
Catalog
Home
Contact Caddock
Corporate
Online Short Form Catalog
Select Product Category
Current Sense Resistors
Power Film Resistors
SMT Resistors
Precision High Voltage Resistors
Low TC and Ultra Precision Resistors
High Performance Resistor Networks
Sudden Fault Fusing Resistor
Download full datasheet (pdf 124K)
Download full datasheet (pdf 132K)

Current Sense Resistors
• Non-Inductive Designs • High Stability • Surface Mount

MP2060 Low Resistance, Clip Mount TO-220 Style Power Film Resistor

Type SR10 and SR20 High Stability Precision Current Sense Resistors

Low Resistance, Lower Cost MP930 TO-220 and MP915 TO-126 Style Power Film Resistors

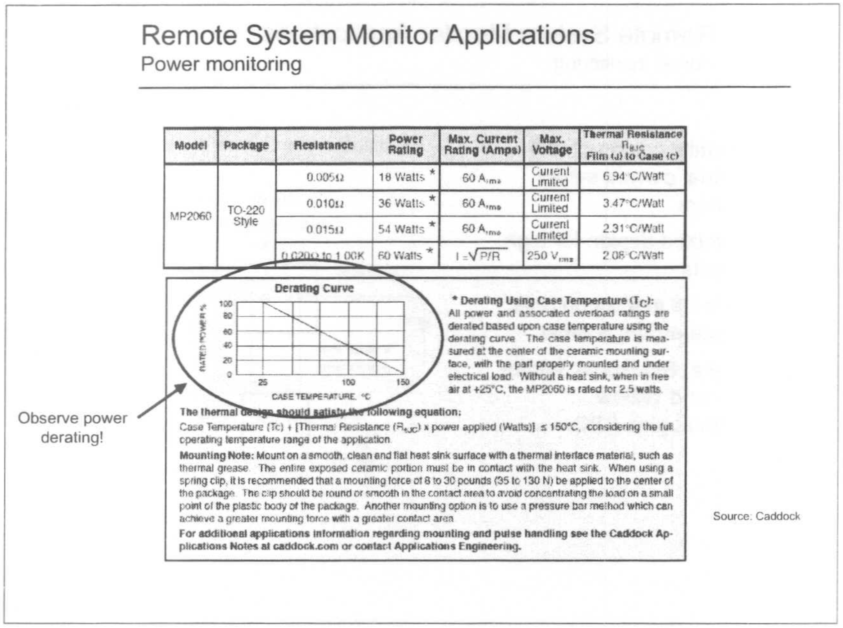
Example of current sense resistors available from CADDOCK

Current Sense Resistors

- Must withstand high current levels
- May dissipate significant power
- May require heat sinking
- Often require values 5mΩ or less
- Sources: Caddock, IRC, Riedon, Ohmite, Willow Technologies

Selection of the shunt resistor boils down to the resistance value, power dissipation requirements, tolerance and physical installation limitations. Shown here is a line of current sense resistors offered by Caddock. The particular power resistor shown, the MP2060, has a 60W rating when attached to a suitable heat sink. It handles currents up to 60 amperes with a resistance as low as 5mΩ.

The resistor tolerance availability depends on the resistance range. For very low values, such as 5mΩ, ± 5% is often the tolerance specification.


Derating Curve
RATED POWER %
100
80
60
40
20
0
25
100
150
CASE TEMPERATURE, °C

Remote System Monitor Applications

Power monitoring

Model	Package	Resistance	Power Rating	Max. Current Rating (Amps)	Max. Voltage	Thermal Resistance $R_{\theta JC}$ Film (J) to Case (c)
MP2060	TO-220 Style	0.005Ω	18 Watts *	60 A_{rms}	Current Limited	6.94°C/Watt
		0.010Ω	36 Watts *	60 A_{rms}	Current Limited	3.47°C/Watt
		0.015Ω	54 Watts *	60 A_{rms}	Current Limited	2.31°C/Watt
		0.020Ω to 1.00K	60 Watts *	$I = \sqrt{P/R}$	250 V_{rms}	2.08°C/Watt

*** Derating Using Case Temperature (T_C):** All power and associated overload ratings are derated based upon case temperature using the derating curve. The case temperature is measured at the center of the ceramic mounting surface, with the part properly mounted and under electrical load. Without a heat sink, when in free air at +25°C, the MP2060 is rated for 2.5 watts.

The thermal design should satisfy the following equation:

Case Temperature (Tc) + [Thermal Resistance ($R_{\theta JC}$) x power applied (Watts)] ≤ 150°C, considering the full operating temperature range of the application.

Mounting Note: Mount on a smooth, clean and flat heat sink surface with a thermal interface material, such as thermal grease. The entire exposed ceramic portion must be in contact with the heat sink. When using a spring clip, it is recommended that a mounting force of 8 to 30 pounds (35 to 130 N) be applied to the center of the package. The clip should be round or smooth in the contact area to avoid concentrating the load on a small point of the plastic body of the package. Another mounting option is to use a pressure bar method which can achieve a greater mounting force with a greater contact area.

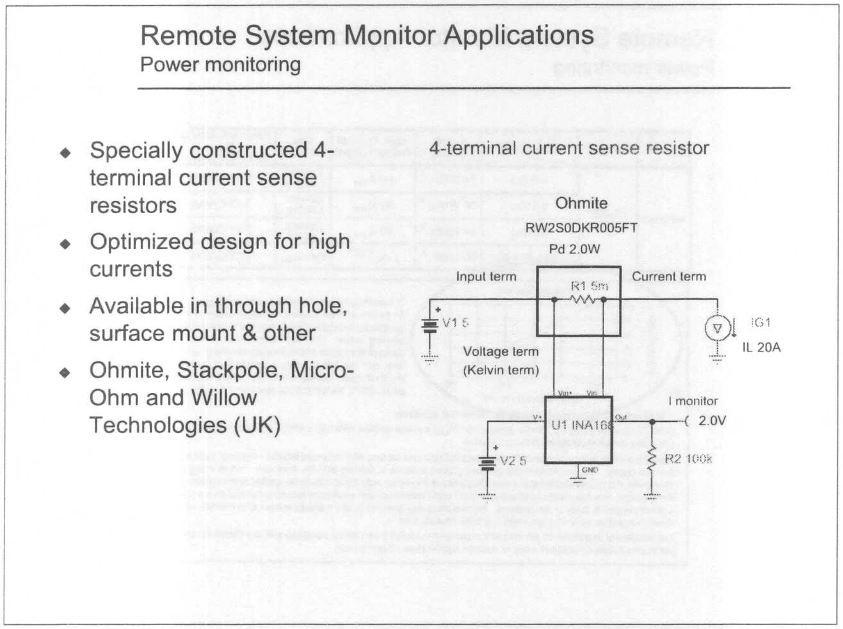
For additional applications information regarding mounting and pulse handling see the Caddock Applications Notes at caddock.com or contact Applications Engineering.

Observe power derating!

Source: Caddock

Even though a sense resistor may have a high power rating, that rating will only apply up to a specific case or ambient temperature; in many cases that temperature is 25°C. Above this the power rating is de-rated falling linearly to zero at some higher temperature. Usually this occurs at 125°C or 150°C.

Therefore, cooling in the form of heat sinking or fan must be employed to reduce the resistor's operating temperature to acceptable levels.



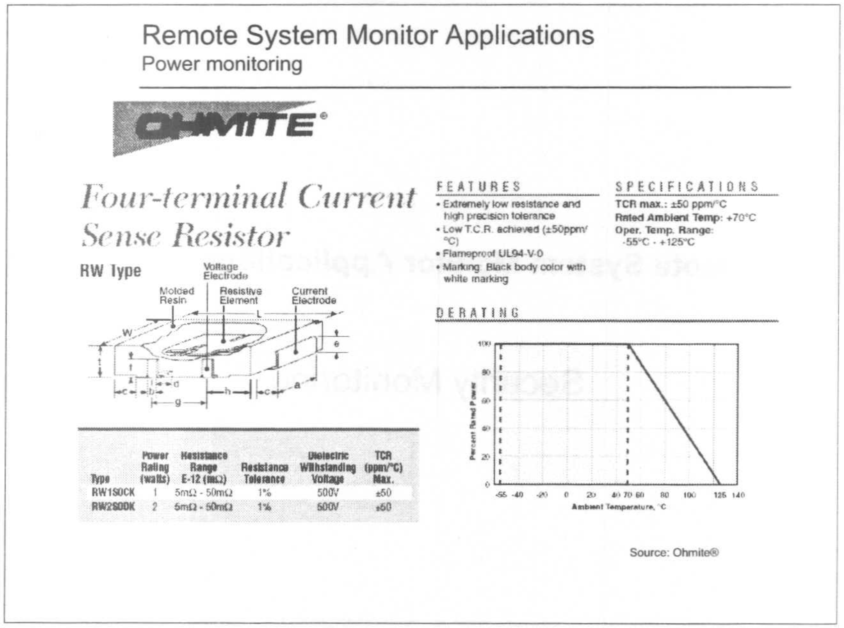
Remote System Monitor Applications

Power monitoring

- Specially constructed 4-terminal current sense resistors
- Optimized design for high currents
- Available in through hole, surface mount & other
- Ohmite, Stackpole, Micro-Ohm and Willow Technologies (UK)

A shunt resistor configuration that is becoming more readily available is the 4-terminal current sense resistor. It not only has the 2 high current "in" and "out" terminals, but two additional that connect internally, directly to the resistive element termination end-points.

These two terminals are then connected to an active sensing circuit such as the INA168. The voltage measured by the INA is nearly devoid of any voltage drop associated with the resistor conductors - the "in" and "out" terminals. This provides an accurate measure of the voltage across the resistor element and not the combined voltages across the resistor element and termination conductors.



Remote System Monitor Applications

Power monitoring

OHMITE®

Four-terminal Current Sense Resistor

RW Type

Voltage Electrode
Molded Resin
Resistive Element
Current Electrode
L
W
t
e
f
d
c
b
g
h
c
a

Type	Power Rating (watts)	Resistance Range E-12 (mΩ)	Resistance Tolerance	Dielectric Withstanding Voltage	TCR (ppm/°C) Max.
RW1S0CK	1	5mΩ - 50mΩ	1%	500V	±50
RW2S0DK	2	5mΩ - 50mΩ	1%	500V	±50

FEATURES

- Extremely low resistance and high precision tolerance
- Low T.C.R. achieved (±50ppm/°C)
- Flameproof UL94-V-0
- Marking: Black body color with white marking

SPECIFICATIONS

TCR max.: ±50 ppm/°C
Rated Ambient Temp: +70°C
Oper. Temp. Range:
-55°C - +125°C

DERATING

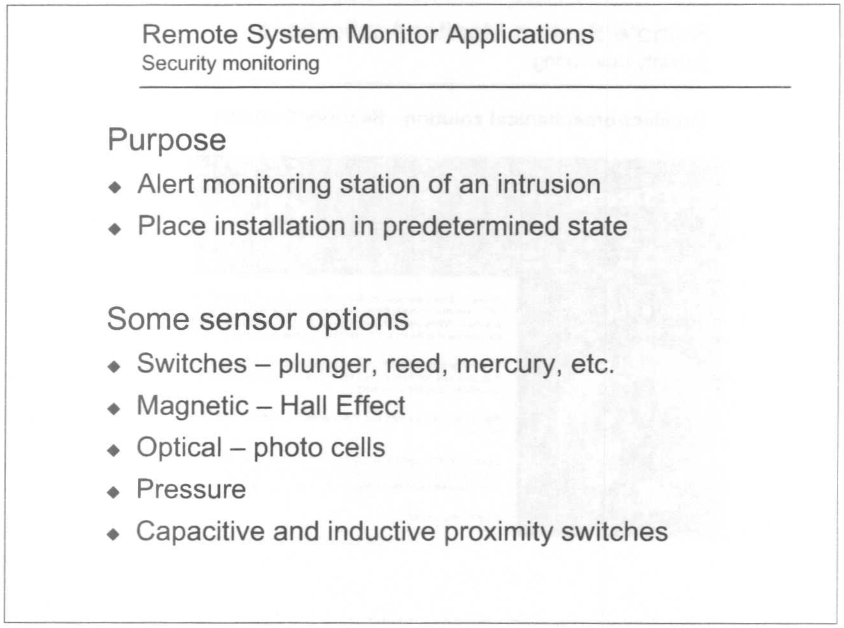
Source: Ohmite®

This excerpt from the Ohmite 4-terminal current sense resistor data sheet shows the construction details of this type of sense resistor. This is a low power resistor rated at 1 or 2 Watts. Notice that the tolerance is rated at ±1%, which is much better than typical ±5% associated with the high power dissipation sense resistors.

Remote System Monitor Applications

Security Monitoring

In the unlikely event that the security of a remote site is breached it is important that the event be detectable as soon as possible, so that a preplanned course of action can be taken to minimize damage to the site equipment and assure the safety of personnel.



Remote System Monitor Applications

Security monitoring

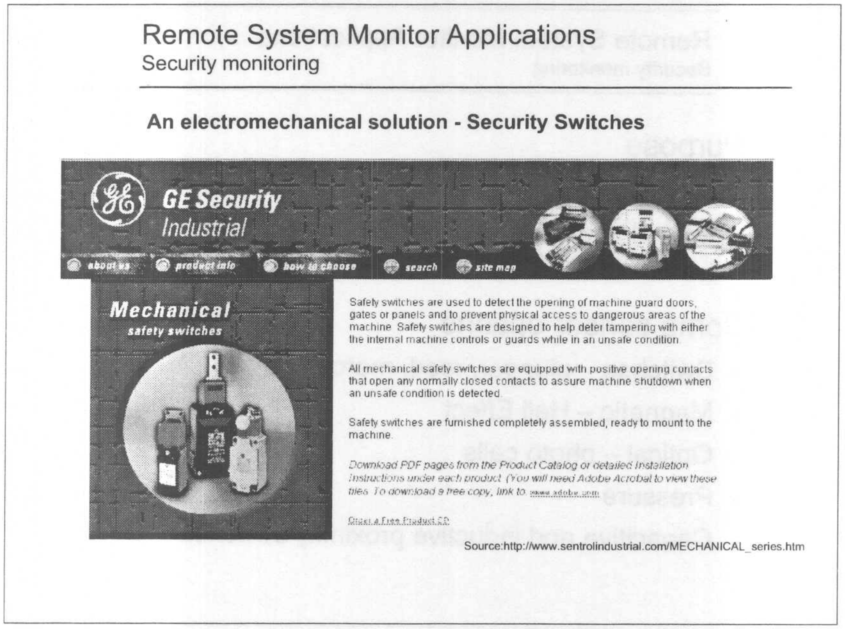
Purpose

- Alert monitoring station of an intrusion
- Place installation in predetermined state

Some sensor options

- Switches – plunger, reed, mercury, etc.
- Magnetic – Hall Effect
- Optical – photo cells
- Pressure
- Capacitive and inductive proximity switches

Various sensor types exist that will help detect the presence of intruders at the site. Most sensors when triggered provide a “1” or “0” response and activate an alarm response. Sometimes it may be desirable to have a more analog response such that one can discriminate between a person and small animal. The response would likely be much different for the two.



Remote System Monitor Applications

Security monitoring

An electromechanical solution - Security Switches

Safety switches are used to detect the opening of machine guard doors, gates or panels and to prevent physical access to dangerous areas of the machine. Safety switches are designed to help deter tampering with either the internal machine controls or guards while in an unsafe condition.

All mechanical safety switches are equipped with positive opening contacts that open any normally closed contacts to assure machine shutdown when an unsafe condition is detected.

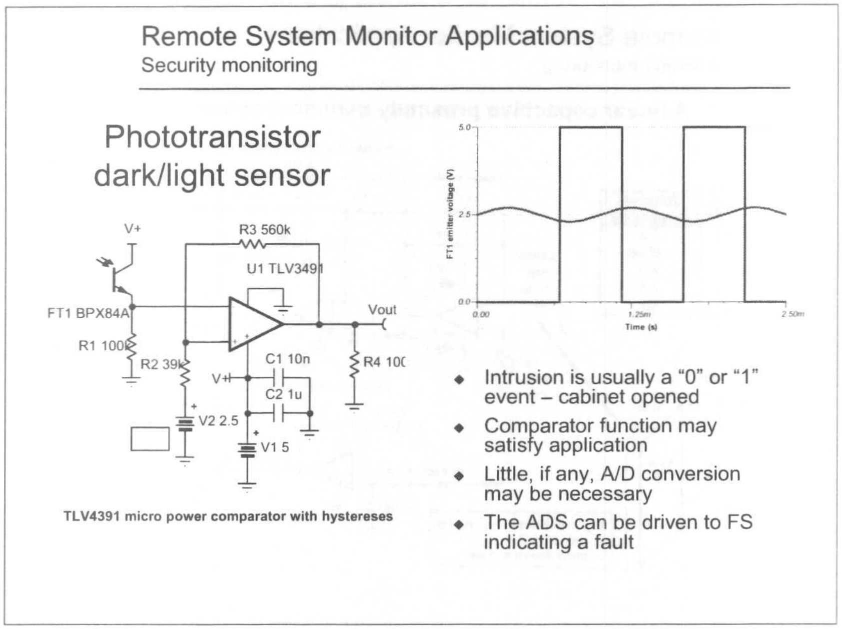
Safety switches are furnished completely assembled, ready to mount to the machine.

Download PDF pages from the Product Catalog or detailed Installation Instructions under each product. (You will need Adobe Acrobat to view these files. To download a free copy, link to www.adobe.com

Order a Free Product CD

Source:http://www.sentrolindustrial.com/MECHANICAL_series.htm

Mechanically actuated switches provide a simple means for detecting a change in state at the remote site. They are often manually actuated plunger type or magnetically actuated reed switches.



Remote System Monitor Applications

Security monitoring

Phototransistor dark/light sensor

TLV4391 micro power comparator with hystereses

- Intrusion is usually a “0” or “1” event – cabinet opened
- Comparator function may satisfy application
- Little, if any, A/D conversion may be necessary
- The ADS can be driven to FS indicating a fault

A photo optical device such as a phototransistor can be used to detect the presence of light such as when a cabinet door is opened, or when the lighting changes at the remote installation.

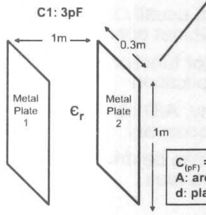
A TLV3491 comparator switches state when the phototransistor conducts subjected to a light source. The comparator has a “1” or “0” output, thus an A-to-D conversion is not required. Hysteresis, added to the comparator circuit, reduces the circuit’s noise sensitivity at the threshold.

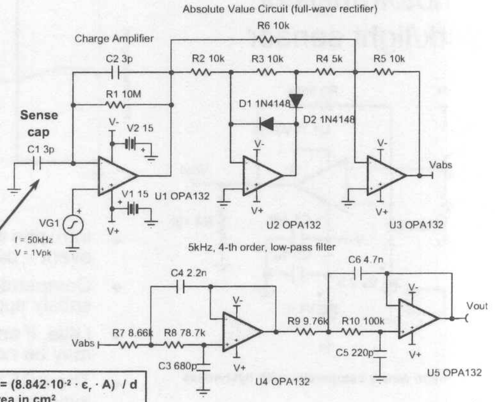
Remote System Monitor Applications

Security monitoring

A linear capacitive proximity sensor proposal

Dielectric	Relative Permittivity
Vacuum	1.0
Air	1.0006
Teflon™	2.0
Glass	7.5
Water	80





$C_{(pF)} = (8.842\cdot10^{-2}\cdot\epsilon_r\cdot A)\ /\ d$

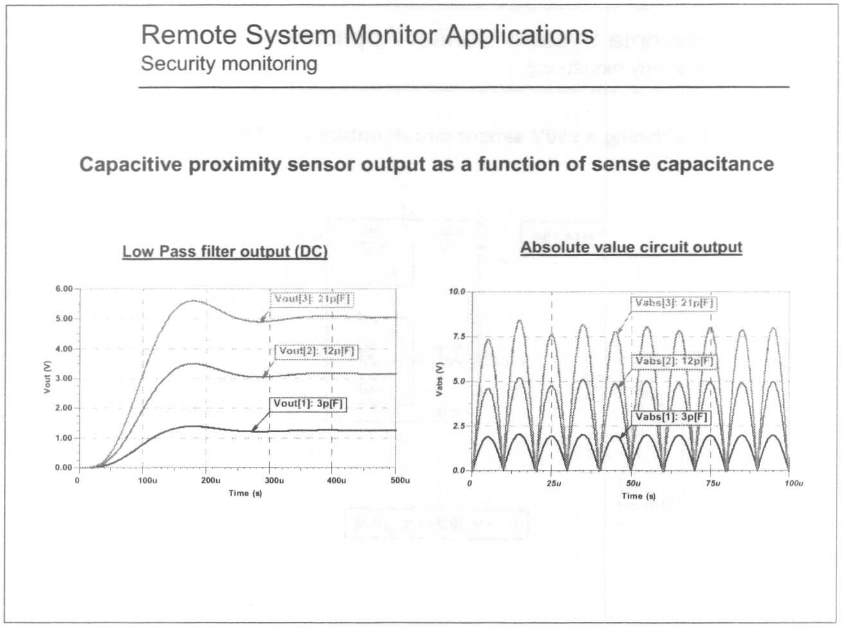
A: area in cm²

d: plate distance in cm

Presented here is a proposal for an analog capacitive proximity sensor. Unlike most conventional capacitive proximity sensors that produce a “1” or “0” output, this sensor produces a DC output that is a function of the size and relative permittivity of an object passing between the sense capacitor plates. Different sized and density objects will produce a different output voltage.

The oscillator input for this circuit is a 50kHz sine source. The low frequency and sine characteristic keep RFI problems at a minimum.

This proximity detector may prove difficult to construct because of the very small capacitances involved, parasitic capacitances in the sensor circuit, and noise pick up. An actual circuit would benefit from larger charge amplifier capacitances, but they may be difficult to fabricate and physically locate. Nonetheless, if this can be overcome then the circuit implementation is viable.



Remote System Monitor Applications

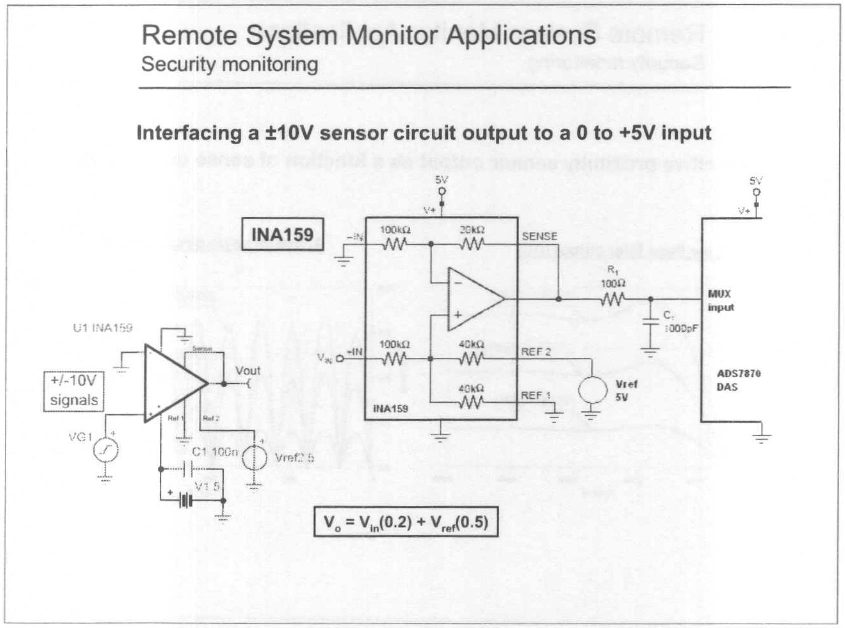
Security monitoring

Capacitive proximity sensor output as a function of sense capacitance

Low Pass filter output (DC)

Absolute value circuit output

These plots show the distinct output levels associated with a change in sensor capacitance. When an object passes between the plates the dielectric's relative permittivity increases, increasing the sense capacitance and gain of the charge amplifier. After rectification by the absolute value stage the pulsating DC voltage is then filtered by the low pass stages. A corresponding DC level appears at the output.



Remote System Monitor Applications

Security monitoring

Interfacing a ±10V sensor circuit output to a 0 to +5V input

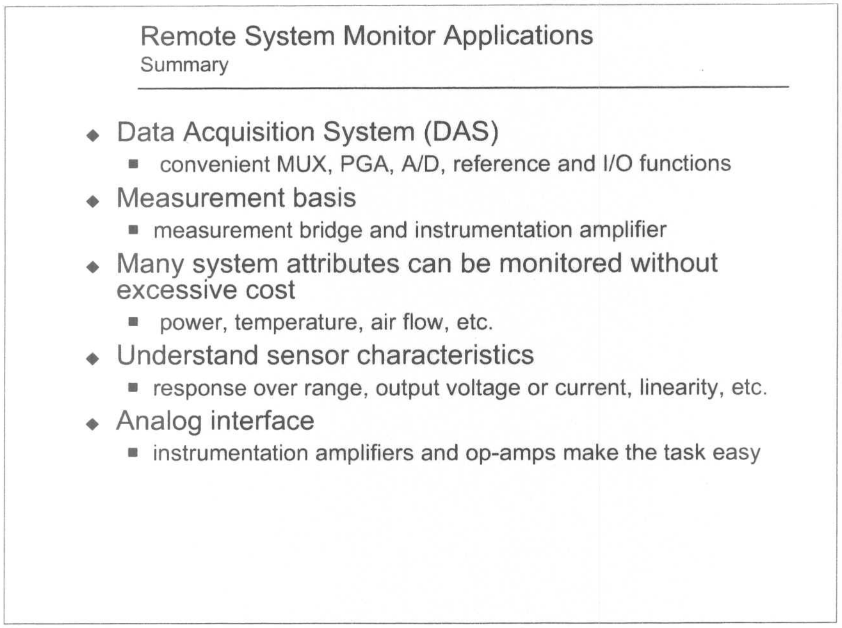
$$\boxed{V_o = V_{in}(0.2) + V_{ref}(0.5)}$$

The output from the capacitive proximity sensor circuit can be large enough to overdrive the DAS input - especially when the dielectric permittivity is very high compared to air. When using sensor circuits that have a high unipolar or bipolar output voltage the INA159 may be the perfect interface between it and the DAS.

The INA159 has the transfer function: $V_o = V_{in}\,(0.2) + V_{ref}\,(0.5)$

Thus, the output signal will be 1/5 the input value plus a DC level equal to ½ the voltage applied to the Ref 2 input. This DC level can be conveniently set to the DAS mid-scale input voltage. If the Ref 2 voltage is set to 5V, then this DC level will be 2.5V, which works nicely with the ADS7870. It is perfect for interfacing high output, such as ±10V ($20V_{p-p}$) sensor outputs to a 0 to 5V input range device.

The INA159 has precision gain scaling and a low voltage offset (±100µV).



Remote System Monitor Applications
Summary

- Data Acquisition System (DAS)
  - convenient MUX, PGA, A/D, reference and I/O functions
- Measurement basis
  - measurement bridge and instrumentation amplifier
- Many system attributes can be monitored without excessive cost
  - power, temperature, air flow, etc.
- Understand sensor characteristics
  - response over range, output voltage or current, linearity, etc.
- Analog interface
  - instrumentation amplifiers and op-amps make the task easy

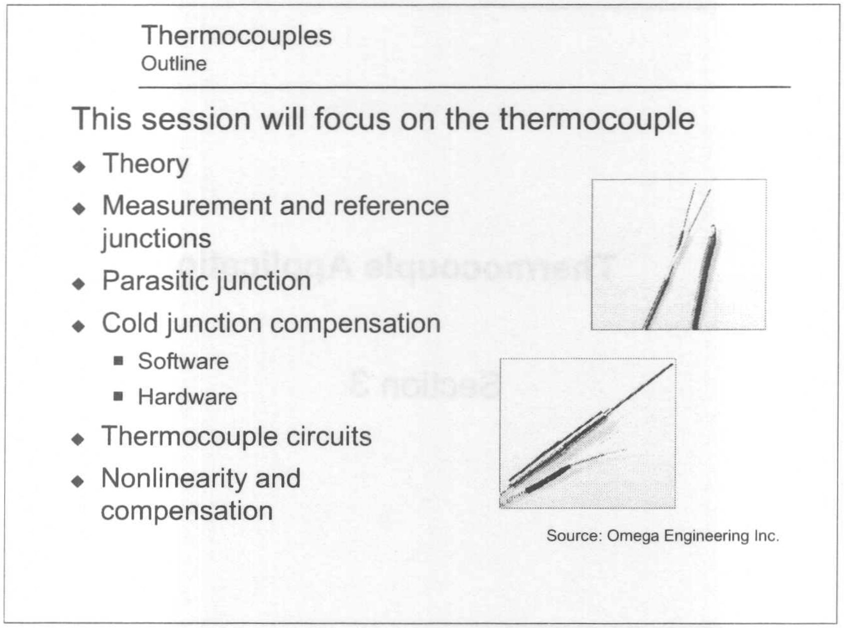
Monitoring a remote system's measurement attributes can be accomplished without adding excessive circuit complexity and cost. An integrated data acquisition system (DAS) makes the collection of sensor outputs, subsequent signal conditioning and data conversion easy.

A measurement bridge, in conjunction with an op-amp or instrumentation amplifier, can provide the basis for a measurement system. The amplifier can be configured to provide the required signal conditioning such as amplification, level shifting and filtering. Since the DAS has a built in amplifier even further voltage gain can be applied to any channel.

The sensors may be fairly simple in design and application as demonstrated in the circuit examples. And as long as their characteristics and limitations are well understood, then they can be properly applied to the measurement task.

Thermocouple Application

Section 3



Thermocouples

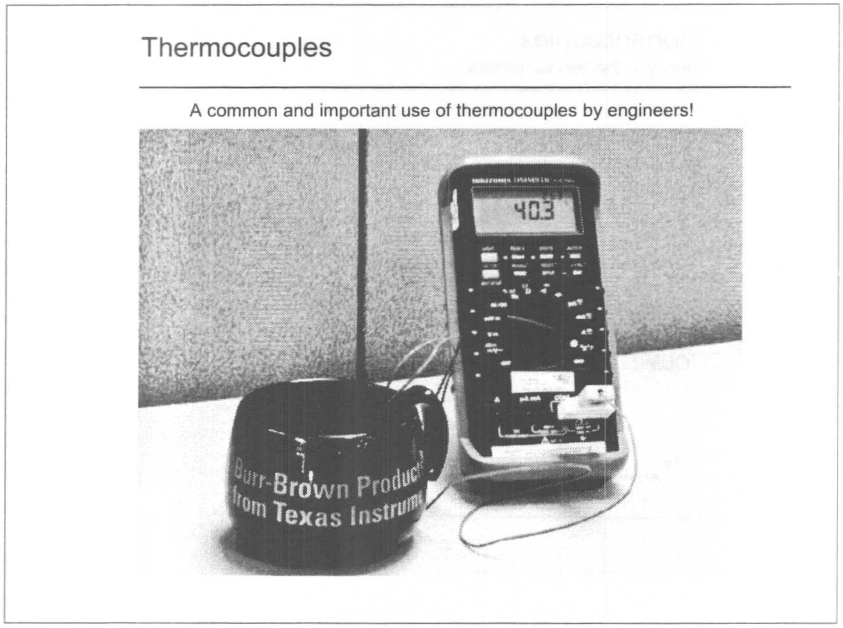
Outline

This session will focus on the thermocouple

- Theory
- Measurement and reference junctions
- Parasitic junction
- Cold junction compensation
  - Software
  - Hardware
- Thermocouple circuits
- Nonlinearity and compensation

Source: Omega Engineering Inc.

Thermocouples are a popular temperature sensor choice due to their wide temperature range capability and rugged design. This session will focus on basic thermocouple theory, principles and how one goes about applying them in a manner such that they produce their best performance.

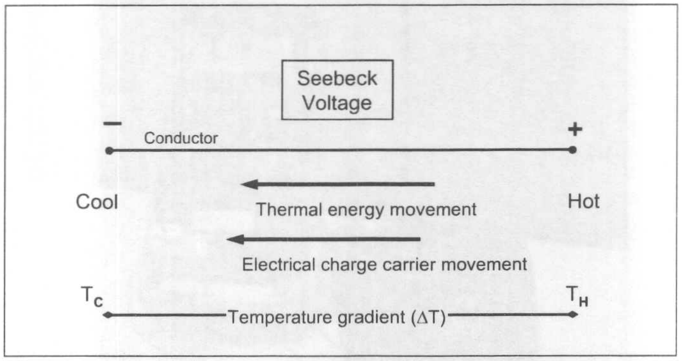


Thermocouples

A common and important use of thermocouples by engineers!

Thermocouples

Theory – the fundamentals



Seebeck Voltage

– Conductor +

Cool

Hot

Thermal energy movement

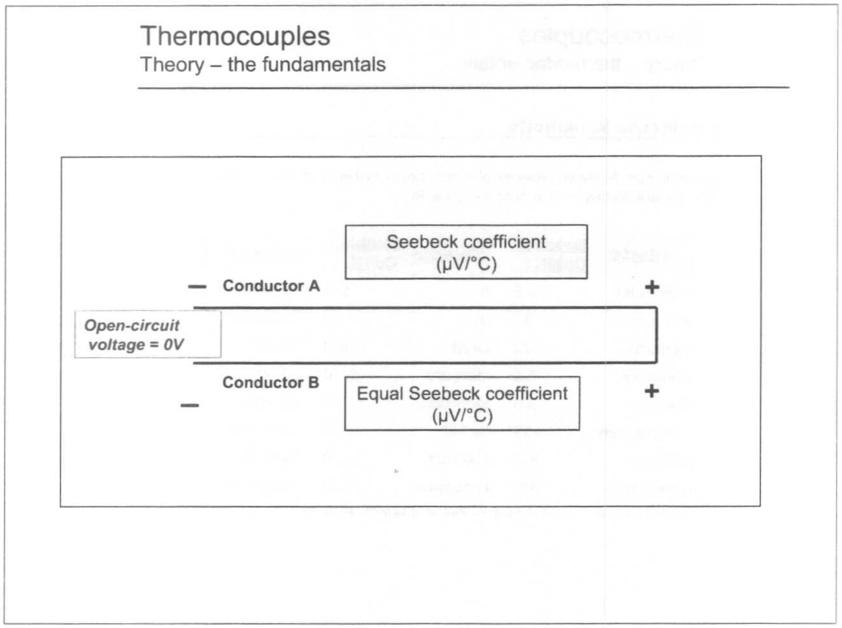
Electrical charge carrier movement

T_C Temperature gradient (ΔT) T_H

A simple wire of any metal will produce a voltage when there is a temperature difference between the two ends. Yes… believe it.

www.dataforth.com/catalog/pdf/an106.pdf

When one end of a conductive material is heated to a temperature larger than the opposite end, the electrons at the hot end are more thermally energized than the electrons at the cooler end. These more energetic electrons begin to diffuse toward the cooler end. Of course, charge neutrality is maintained; however, this redistribution of electrons creates a negative charge at the cool end and an equal positive charge (absence of electrons) at the hot end. Consequently, heating one end of a conductor creates an electrostatic voltage due to the redistribution of thermally energized electrons throughout the entire material. This is referred to as the “Seebeck effect.” While a single wire does not form a thermocouple, this “Seebeck effect” is the fundamental property that governs thermocouple operation.



Thermocouples

Theory – the fundamentals

Seebeck coefficient (µV/°C)

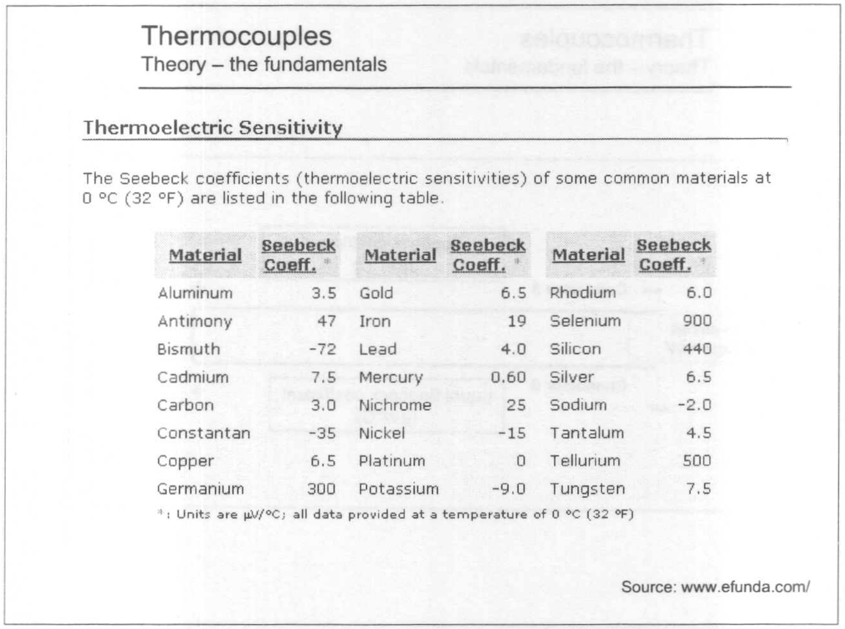
– **Conductor A** +

Open-circuit voltage = 0V

– **Conductor B** +

Equal Seebeck coefficient (µV/°C)

Direct measurement of the Seebeck voltage of a single wire is impossible. Another wire of the same metal produces an identical Seebeck voltage resulting in a net voltage of 0V at the measurement points.



Thermocouples

Theory – the fundamentals

Thermoelectric Sensitivity

The Seebeck coefficients (thermoelectric sensitivities) of some common materials at 0 °C (32 °F) are listed in the following table.

Material	Seebeck Coeff. *	Material	Seebeck Coeff. *	Material	Seebeck Coeff. *
Aluminum	3.5	Gold	6.5	Rhodium	6.0
Antimony	47	Iron	19	Selenium	900
Bismuth	-72	Lead	4.0	Silicon	440
Cadmium	7.5	Mercury	0.60	Silver	6.5
Carbon	3.0	Nichrome	25	Sodium	-2.0
Constantan	-35	Nickel	-15	Tantalum	4.5
Copper	6.5	Platinum	0	Tellurium	500
Germanium	300	Potassium	-9.0	Tungsten	7.5

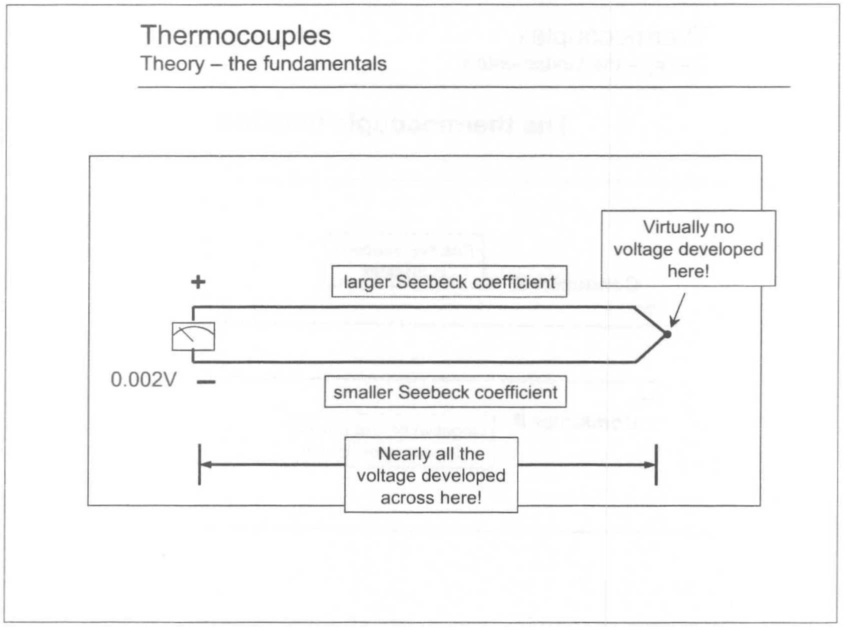
*: Units are µV/°C; all data provided at a temperature of 0 °C (32 °F)

Source: www.efunda.com/

Different metals, metal alloys and semiconductor materials are employed in the construction of thermocouples. Their thermoelectric sensitivities, or Seebeck coefficients, can vary significantly in magnitude and may be positive or negative.

The materials listed have been well characterized, standardized, and form the basis for the commonly available thermocouples.

Note that different tables may list a somewhat different Seebeck coefficient for a given material. Be sure to note the temperature at which the coefficient is specified. Thermocouples are not perfectly linear across temperature. They may produce a different Seebeck voltage coefficient within the different temperature ranges that they operate. This occurs because the Seebeck voltage generated is dependent on a complex mix consisting of the Seebeck, Peltier and Thomson effects.



Thermocouples

Theory – the fundamentals

Virtually no voltage developed here!

+

larger Seebeck coefficient

0.002V −

smaller Seebeck coefficient

Nearly all the voltage developed across here!

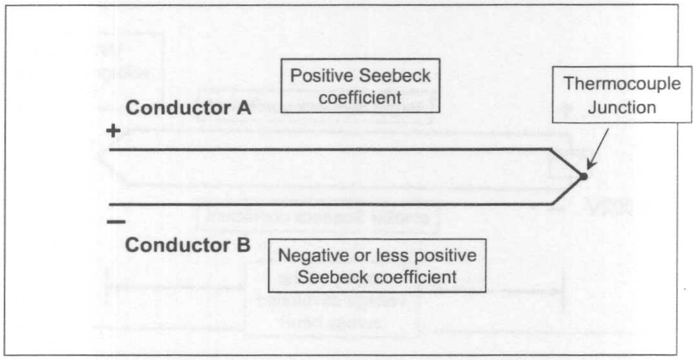
Perhaps the most misunderstood issue regarding thermocouples is that no voltage is produced at the measurement junction. The junction completes the circuit so that current flow can take place. A voltage is developed along each wire as the temperature changes. The voltage difference is observed at the receiving end because the two differing metals have different Seebeck coefficients and produce a voltage difference at the meter point.

Misinformation about thermocouples abounds on the internet with statements such as "… the junction between two metals generates a voltage which is a function of temperature." Many other references and web sites make the same error. A more accurate explanation can be found at:

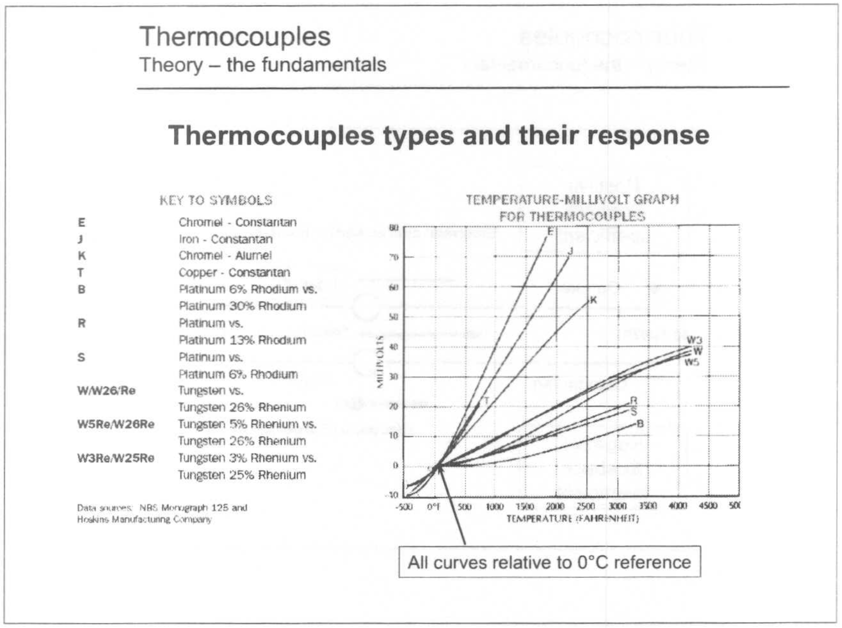
www.dataforth.com/catalog/pdf/an106.pdf

Thermocouples
Theory – the fundamentals

The thermocouple junction


Positive Seebeck coefficient
Thermocouple Junction
Conductor A
+
–
Conductor B
Negative or less positive Seebeck coefficient

A thermocouple junction is formed when two dissimilar metals, metal alloys or semiconductor materials are joined together. However, the practical thermocouple not only consists of the junction, but connecting leads made of the same dissimilar metals. In use, the thermocouple junction is exposed to the "hot" (or cold) temperature point. The leads connect between the junction and a measurement device located at a different temperature such as room. It is along these lead lengths where the temperature gradient is present resulting in the generation of the two individual Seebeck EMF's.



Thermocouples

Theory – the fundamentals

Thermocouples types and their response

	KEY TO SYMBOLS
E	Chromel - Constantan
J	Iron - Constantan
K	Chromel - Alumel
T	Copper - Constantan
B	Platinum 6% Rhodium vs. Platinum 30% Rhodium
R	Platinum vs. Platinum 13% Rhodium
S	Platinum vs. Platinum 6% Rhodium
W/W26/Re	Tungsten vs. Tungsten 26% Rhenium
W5Re/W26Re	Tungsten 5% Rhenium vs. Tungsten 26% Rhenium
W3Re/W25Re	Tungsten 3% Rhenium vs. Tungsten 25% Rhenium

Data sources: NBS Monograph 125 and Hoskins Manufacturing Company

TEMPERATURE-MILLIVOLT GRAPH FOR THERMOCOUPLES

All curves relative to 0°C reference

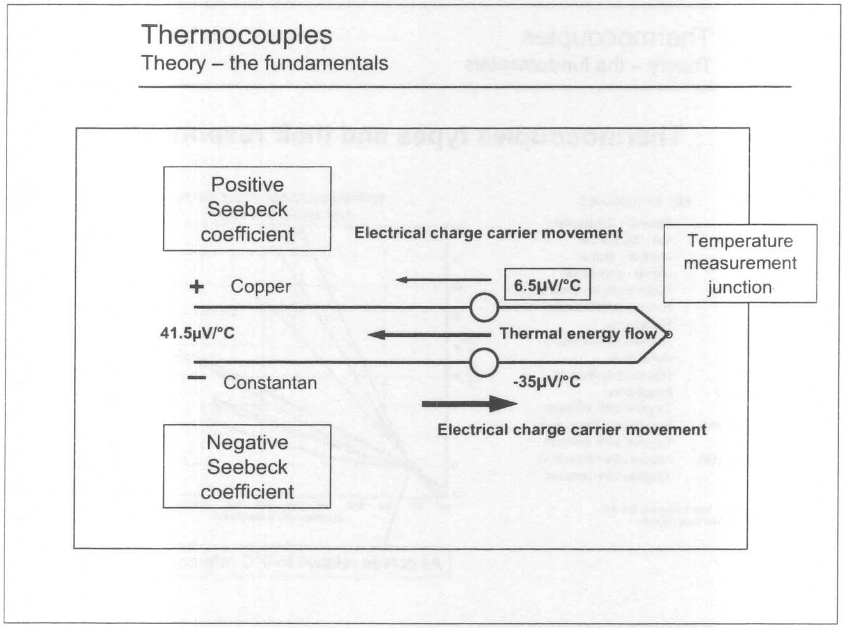
Thermocouples are classified by type which is associated with their useable temperature range, sensitivity and accuracy. The commonly used metals include: chromium, copper, nickel, iron, platinum, rhodium, and rhenium.

This chart provides the thermal response of several different types of thermocouples. Notice that the copper-constantan “type T” thermocouple has a limited use temperature range compared to the others.

Also note the differences in the thermocouple sensitivities and their linearity ΔV/ΔT. Those having a more limited temperature range tend to have better linearity characteristics. Because of poor linearity some higher temperature thermocouples aren’t intended for measuring temperatures below 0°F (-18°C).

As previously mentioned the Seebeck coefficient may be listed with a different value, which may depend on the source of the information. The specified temperature was mentioned as a cause for the difference. For example, the copper-constantan “type-T” thermocouple is listed with a Seebeck coefficient of 41µV/°C at 25°C*, and 38.75µV/°C at 0°C, in the Agilent Technologies, Application note 290. A value of 38µV/°C is often listed.

NOTE: A similar value is given at the “efunda” website which lists a “type-T” Seebeck coefficient of 40.6µV/°C at 25°C.



Thermocouples

Theory – the fundamentals

Positive Seebeck coefficient

Electrical charge carrier movement

Temperature measurement junction

+ Copper

6.5µV/°C

41.5µV/°C

Thermal energy flow

– Constantan

-35µV/°C

Electrical charge carrier movement

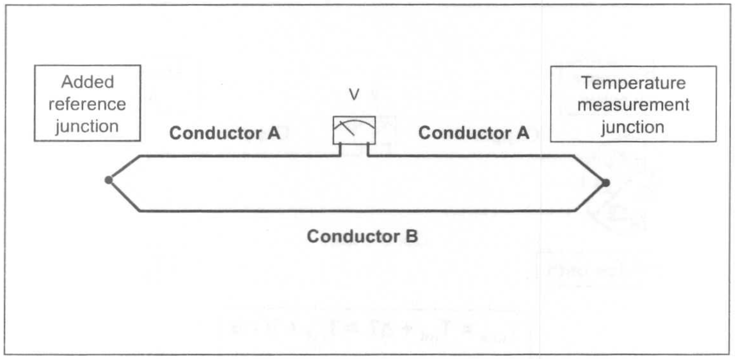
Negative Seebeck coefficient

Different thermocouple materials have different capacities for moving charge carriers in response to thermal flow. The current level in one conductor will overcome or complement the potential for thermally generated current flow in the other conductor. The result is a continuous current flow that is the difference between the currents generated in the two conductors.

For this example, the two selected metals are copper and constantan which have Seebeck coefficients of approximately +6.5µV/°C and -35µV/°C, respectively. The difference between these two coefficients results in a thermocouple sensitivity of about +41.5µV/°C at 0°C.

Thermocouples

Reference junctions



Added reference junction

V

Conductor A

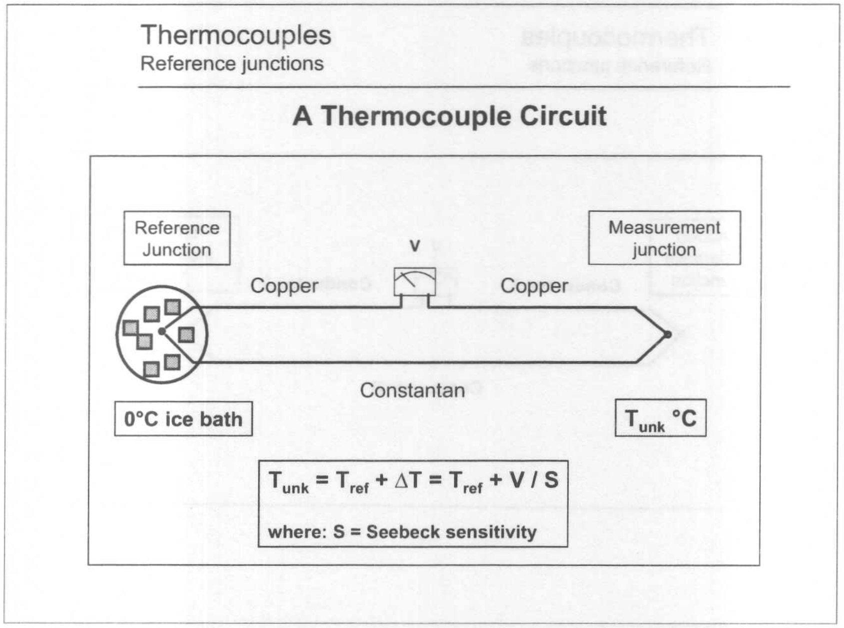
Conductor A

Temperature measurement junction

Conductor B

The thermocouple example in the previous slide had a thermoelectric sensitivity of about 41.5µV/°C. That is an important bit of information, but equally important and missing is a temperature reference point. A temperature change can be measured, but the actual temperature is still an unknown. Adding a second junction and holding it at a known reference temperature allows an unknown temperature at the other junction to be found.

Since the circuit is a continuous loop in which current flows it can be opened and a meter inserted. The voltmeter has a high internal resistance and produces a voltage proportional to the current. Keep in mind that the voltage is strictly dependent upon temperature; the relationship between Seebeck voltage and temperature is fixed. However, the relationship between temperature and current is variable and will depend on the overall circuit resistance.



Thermocouples

Reference junctions

A Thermocouple Circuit

$$T_{unk} = T_{ref} + \Delta T = T_{ref} + V / S$$

where: S = Seebeck sensitivity

Placing the reference junction in an ice bath with a temperature very close to 0°C allows for the unknown temperature to be determined using the following relations:

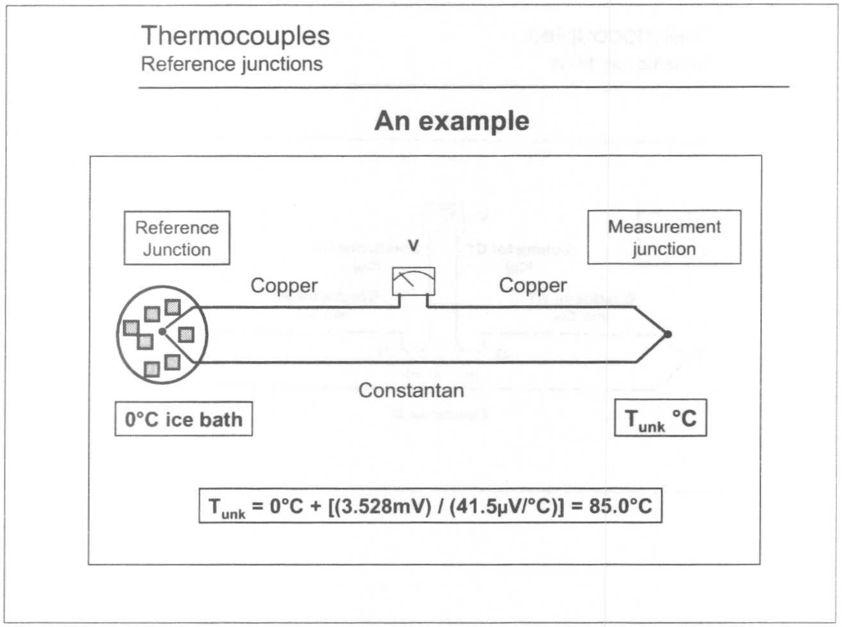
$$V = S \cdot \Delta T$$

$$\Delta T = V / S$$

where: V = measured voltage, S = Seebeck coefficient (V/°C)

Then:

$$T_{unk} = T_{ref} + \Delta T = T_{ref} + V / S$$



Thermocouples
Reference junctions

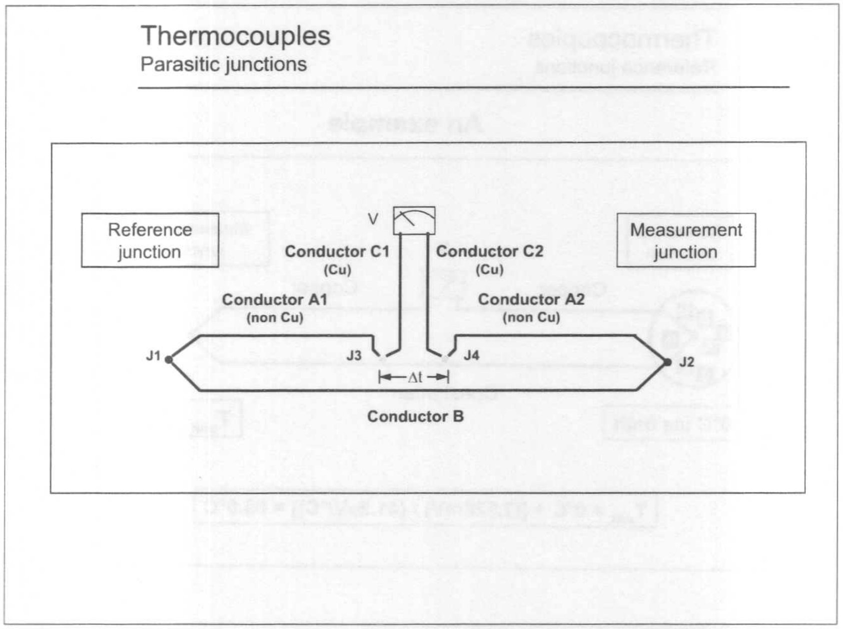
An example

$T_{unk} = 0°C + [(3.528mV) / (41.5µV/°C)] = 85.0°C$

For example:

If a copper-constantan thermocouple produces a voltage of 3.528mV

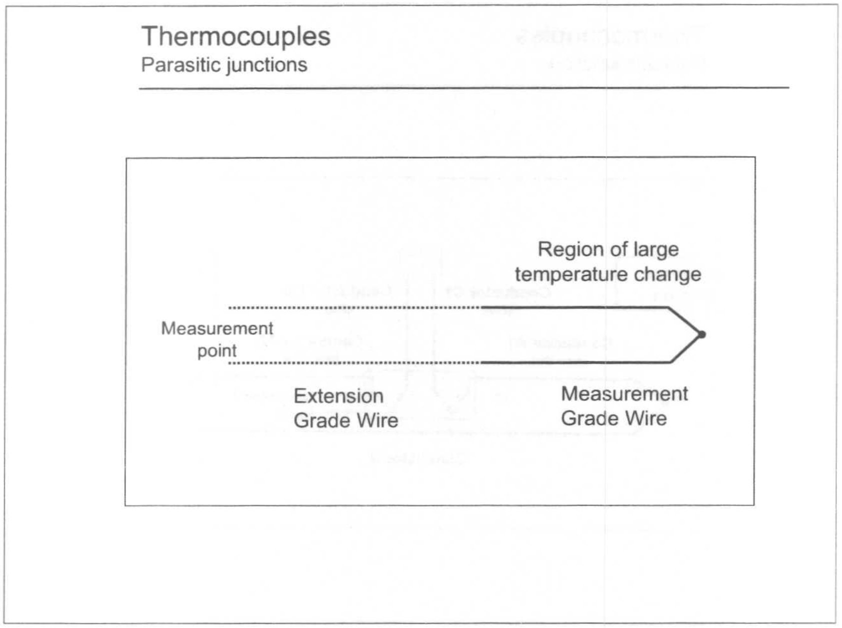
then, $T_{unk} = 0°C + [(3.528mV) / (41.5µV/°C)] = 85.0°C$



Thermocouples

Parasitic junctions

A copper-to-copper connection is unique to the case of the “type T” thermocouple. But when a thermocouple other than the “type T” is employed parasitic thermocouples are created at the meter connections or leads leading to the meter function. These parasitic thermocouples may introduce measurement errors. Each generates a Seebeck voltage dependent on the junction materials and relevant temperature gradient.



Thermocouples

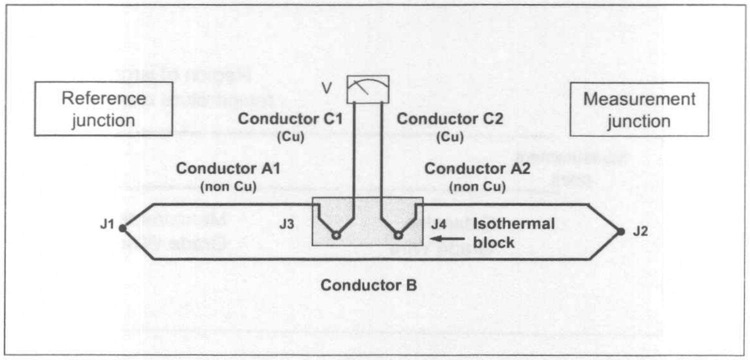
Parasitic junctions

One way to avoid the problems associated with creating parasitic thermocouple junctions is to use extension wires similar in characteristics to the actual thermocouple section.

Thermocouple wire can be relatively expensive and comes in various accuracy grades. Measurement-grade wire is made of higher purity metals and more accurately controlled alloys, thus providing greater accuracy. This higher quality wire is often used only in the region of greatest temperature change where virtually all the voltage is produced. Depending on the application, this may be only in the first few centimeters near the measurement junction. Lower quality wire called "extension grade" can be used to connect to the measurement system without seriously degrading accuracy.

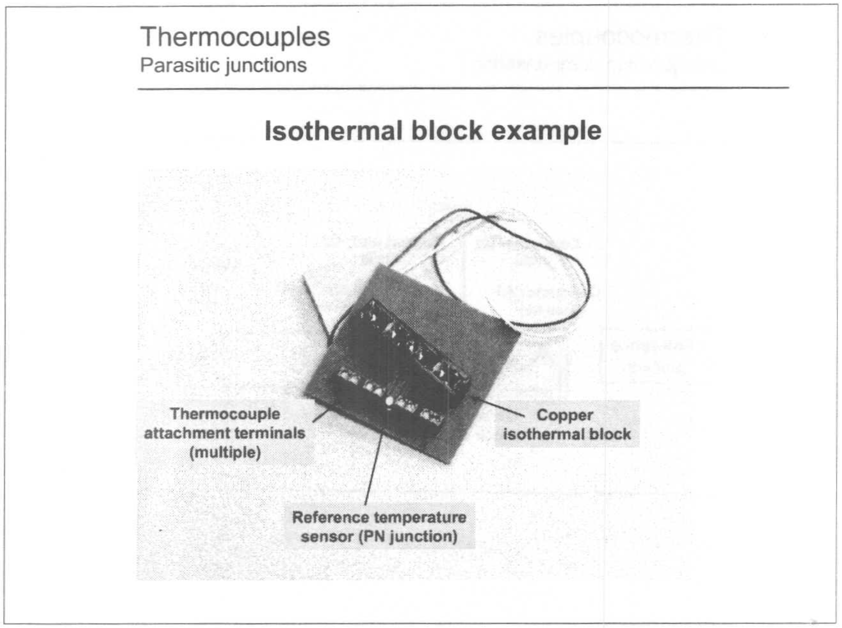
Thermocouples

Parasitic junctions



With the “type T” thermocouple the copper line can be opened and directly connected to copper extension lines without forming parasitic thermocouples. But with other materials that won’t be the case. Even then it’s not the end of the world because the two parasitic junctions, J3 and J4, will produce equal and opposite voltages - provided they are identical and at the same temperature. Moderately accurate measurements will be obtained even if they aren’t.

A way to help assure this is to make the extension wire connections at an isothermal block. The block maintains the two junctions at the same temperature and provides nearly identical electrical connection characteristics. The block must be insulated for the electrical connections and provide for good thermal conductivity between them.



Thermocouples

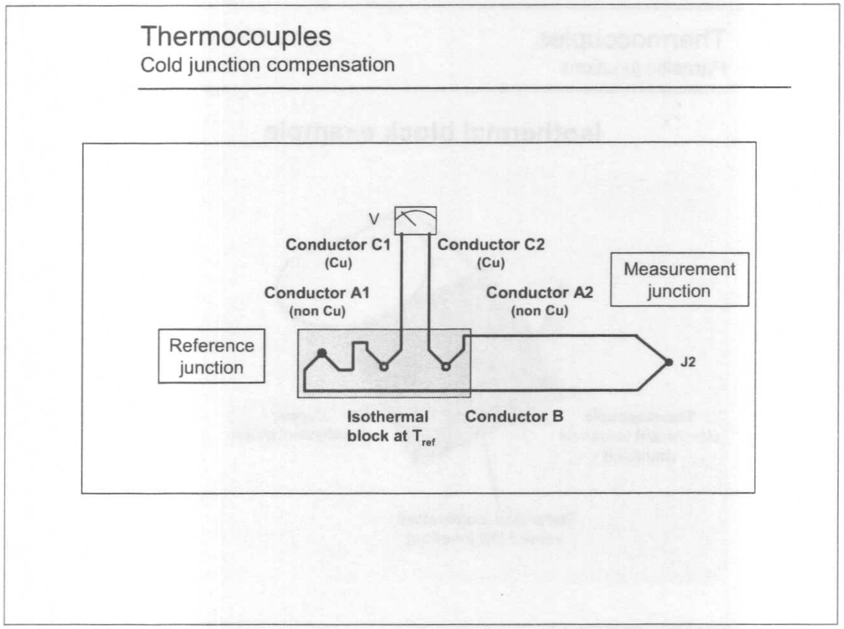
Parasitic junctions

Isothermal block example

Thermocouple attachment terminals (multiple)
Copper isothermal block
Reference temperature sensor (PN junction)

This is an image of an isothermal block that is intended for 4 individual thermocouples. The copper isothermal block fits over plastic terminal blocks. It has sufficient thermal mass such that all of the terminals should be held very close in temperature.

It also has holes along the front edge for the thermocouple wires to pass through and holes on the top to access the terminal block screws.

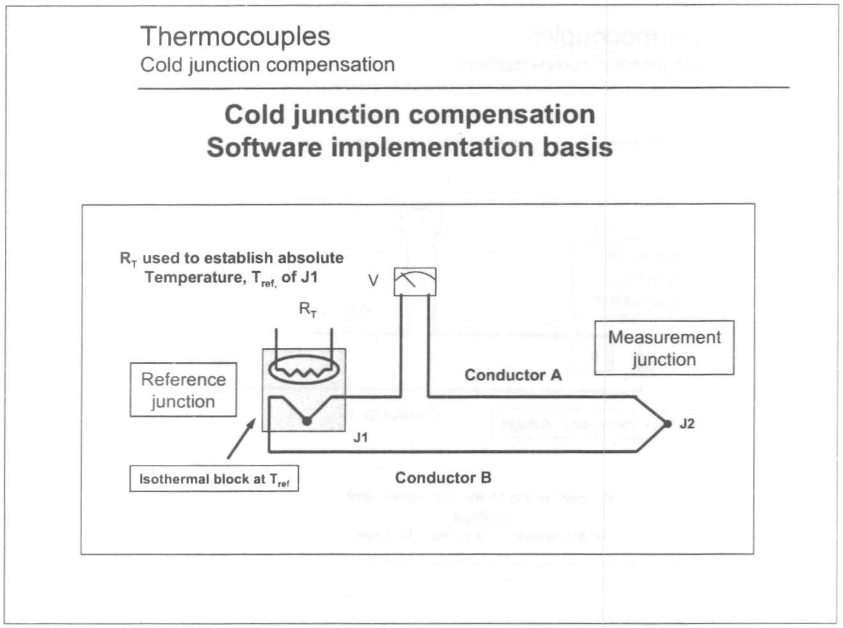


Thermocouples

Cold junction compensation

Often it is not practical to include an ice bath reference as part of the measurement system. Shown here the reference junction has now been located at the isothermal block along with the parasitic junctions. As long as the parasitic junctions are held at a common temperature they will cancel each other's Seebeck voltage contribution.

The reference junction will still require establishment of a reference temperature, but this can be accomplished by software or hardware compensation techniques.



Thermocouples

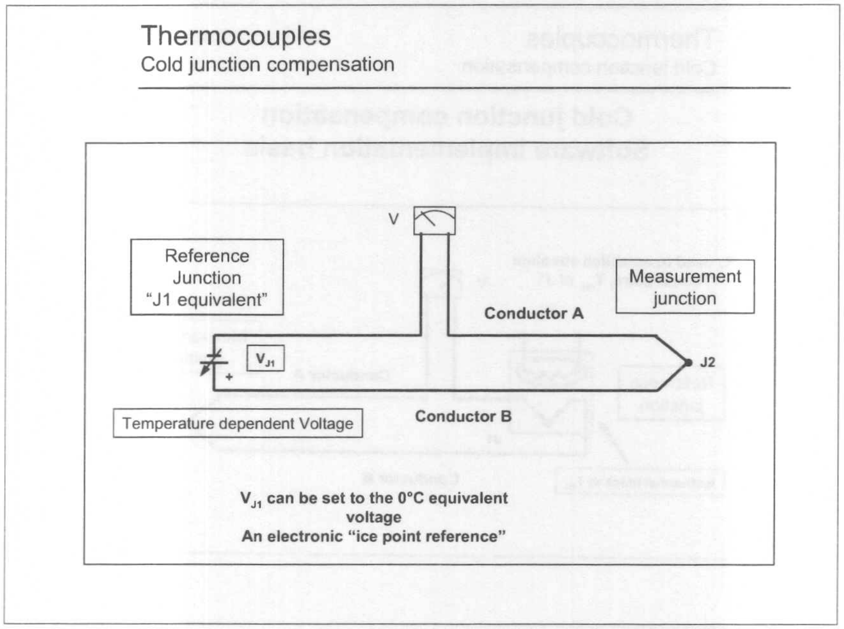
Cold junction compensation

Cold junction compensation
Software implementation basis

R_T used to establish absolute Temperature, $T_{ref,}$ of J1
V
R_T
Measurement junction
Reference junction
Conductor A
J2
J1
Isothermal block at T_{ref}
Conductor B

A secondary temperature sensing transducer such as a thermistor, RTD or semiconductor junction may be attached at the isothermal block to indicate the block's temperature. R_T has a resistance that is proportional to the isothermal blocks temperature. The temperature response characteristics of this secondary transducer must be an established known in order to be utilized. The resistance is then converted to another electrical property such as voltage, and then to its digital equivalent. This compensation voltage can than be summed with the measured voltage in the software. This technique is known as software compensation.

One may question why one wouldn't use this reference transducer to measure the temperature in the first place? The answer is that transducers of this type have a limited useful temperature range when compared to a thermocouple. And they also lack the physical properties required for many high temperature and/or physically demanding applications. Thermocouples are rugged, high temperature transducers that are often subjected to harsh environments with conditions that far exceed what the other transducers can withstand.



Thermocouples

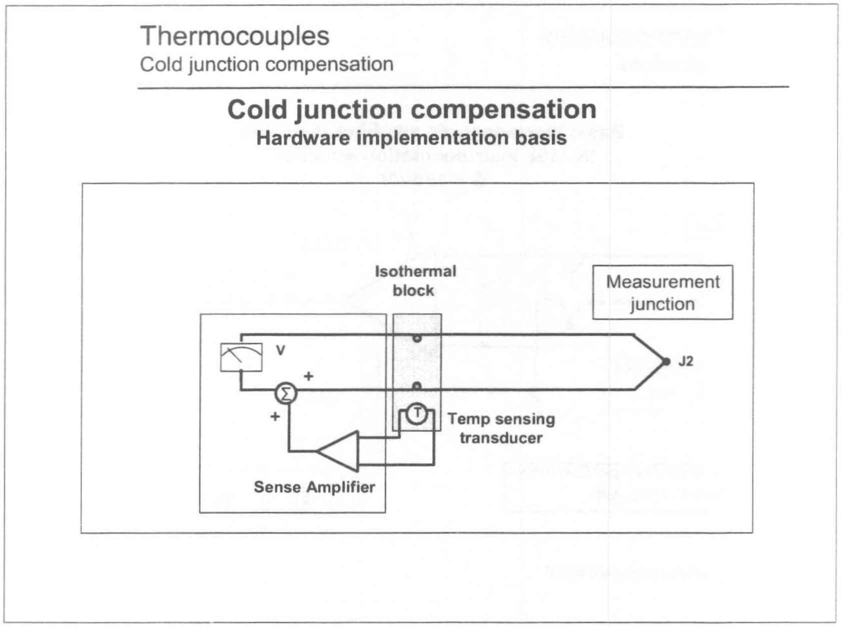
Cold junction compensation

V
Reference Junction "J1 equivalent"
Measurement junction
Conductor A
V_{J1}
+
J2
Temperature dependent Voltage
Conductor B

V_{J1} can be set to the 0°C equivalent voltage

An electronic "ice point reference"

When subjected to an ice bath the reference junction develops a voltage specific to 0°C. An equivalent voltage source can be substituted in place of the junction to serve as a 0°C voltage reference. This electronic substitution for the ice bath is referred to as an "electronic ice point reference." This standard voltage is dependent on the particular thermocouple type and the values are established by the NIST. Electronic ice point references are available for many different types of thermocouples.



Thermocouples
Cold junction compensation

Cold junction compensation

Hardware implementation basis

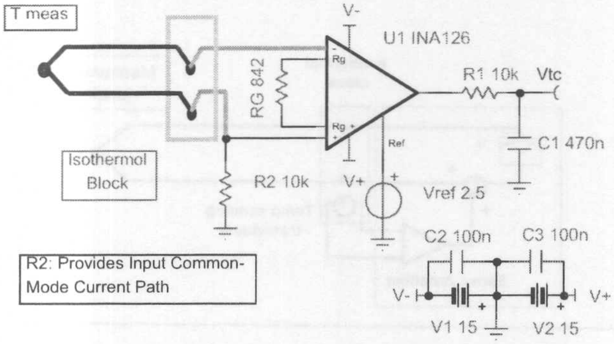
In a practical hardware compensation scheme the secondary transducer's voltage is appropriately gained and summed within the measurement circuit's path. The secondary temperature sensing transducer is mounted to the isothermal block. This can be a thermistor, RTD etc. Its resistance tracks the temperature of the isothermal block and is converted by the sense amplifier to a voltage that is summed or subtracted at the summing junction.

The secondary sense transducer response over temperature has to be taken into account so that the correct voltage is summed into the measurement path.

Thermocouples

Applications

Basic thermocouple amplifier featuring INA126 instrumentation amplifier
G = 100V/V



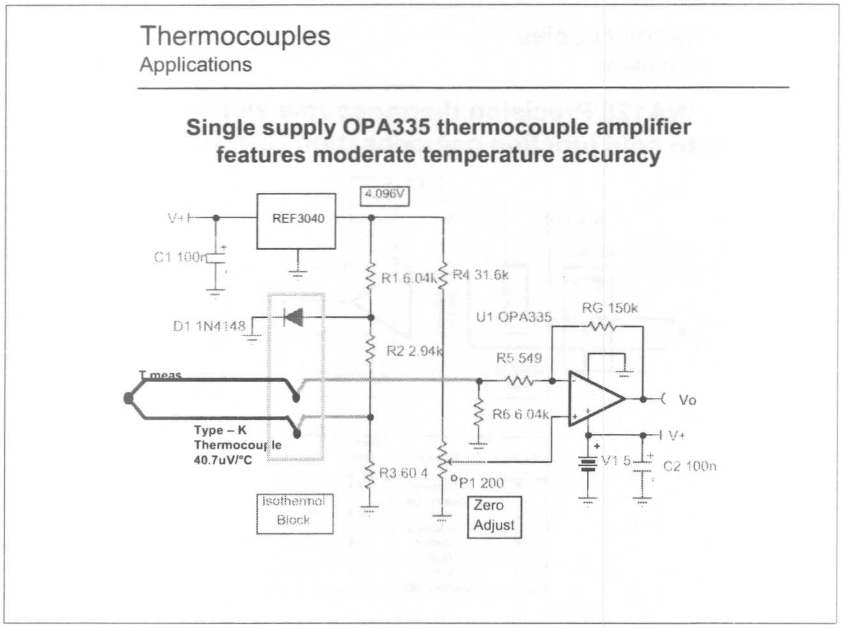
R2: Provides Input Common-Mode Current Path

NOTE: no cold junction compensation!

Since thermocouples produce DC signal levels in the tens or hundreds of microvolts it is necessary to provide additional gain for further signal processing. Interfacing the thermocouple is a simple matter of using a 3-amplifier, instrumentation amplifier. In this case an INA126 MicroPOWER instrumentation amplifier is employed and provides a voltage gain of 100V/V. Despite its very low power usage (Iq = 200µA max) its speed is completely adequate for this type of application. Note that this simple circuit does not include a reference, or equivalent, and only temperature change would be observed. The other complexities can be added to suit the application.

It should be noted that with amplifiers like the INA126 that have extremely high input impedance (≈10^9) that a path must be provided for the input bias currents. With floating transducers, like the thermocouple, this is easily accomplished by adding a resistor off one side to ground (R_2).

One might be tempted to think that this circuit is not useable in its present state; however, it may be suitable for low accuracy applications. The main drawback is the lack of cold junction compensation, but may only introduce a small error if the temperatures being measured are high. For example, with measurement temperatures in excess of 1000°C, the error caused by not including the cold junction temperature would likely be tolerable.



Thermocouples
Applications

Single supply OPA335 thermocouple amplifier features moderate temperature accuracy

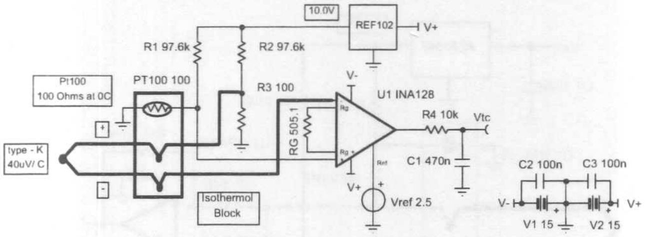
This is a complete thermocouple amplifier for a type K thermocouple. It features an OPA335 CMOS, zero-drift op-amp and includes cold junction compensation (isothermal block) and incorporates a diode thermal sensing circuit for hardware compensation.

This circuit will produce moderately accurate results limited somewhat by the inexact diode characteristics. Although a PN junction is the most linear of all temperature sensors, its accuracy at a given temperature can vary due to the diode's saturation current characteristics. A 10:1 difference in the diode saturation current results in a 60mV difference in forward junction voltage. From one batch of diodes to the next, the forward voltage can be quite different which would result in a different cold junction temperature.

Thermocouples

Applications

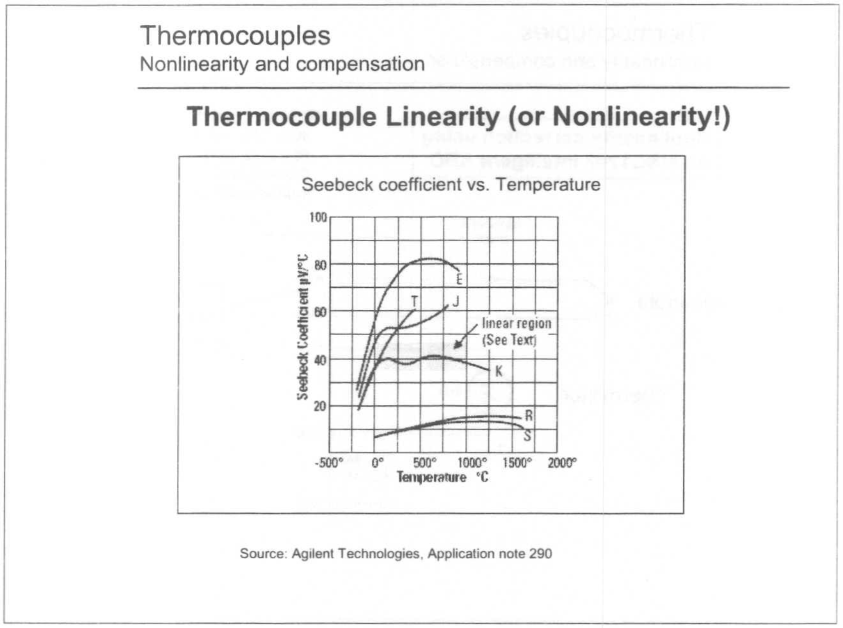
INA128 Precision thermocouple amplifier with cold junction compensation (G = 100V/V)



ISA Type	Material	Seebeck Coeff uV/°C	R1, R2
E	+ Chromel - Constantan	58.5	66.5kΩ
J	+ Iron - Constantan	50.2	76.8kΩ
K	+ Chromel - Alumel	39.4	97.6kΩ
T	+ Copper - Constantan	38.0	102kΩ

This thermocouple amplifier uses the INA128 precision instrumentation amplifier in a gain of 100V/V. Cold junction compensation is accomplished with a Pt100 RTD. It exhibits very good linearity over most of its operating range and the accuracy can be specified with a fraction of a degree. Therefore, from one RTD batch to the next, the temperature accuracy performance can be duplicated.

The table lists the resistor values for R1 and R2 associated with various thermocouples. These resistors establish RTD bias such that the associated voltage corresponds to the block temperature.



Thermocouples

Nonlinearity and compensation

Thermocouple Linearity (or Nonlinearity!)

Seebeck coefficient vs. Temperature

Source: Agilent Technologies, Application note 290

Up to this point we have been using a fixed constant for the Seebeck coefficient, but mention has been made that it will vary within the thermocouple's useable temperature range. For some types of thermocouples the coefficient may be 2 to 3 times higher within portions of the operating temperature range. This lack of linearity, or nonlinearity, will result in large temperature measurement errors if some form of linearization is not applied.

There are a number of ways one may go about correcting for the thermocouple's nonlinearity, but all rely on applying linearization coefficients to the measured voltage. The coefficients are often mathematically derived or acquired from look-up tables. Categorization and fast algorithms can be used to speed up the process. The choice really depends on the power of the data acquisition system employed in the measurement system.

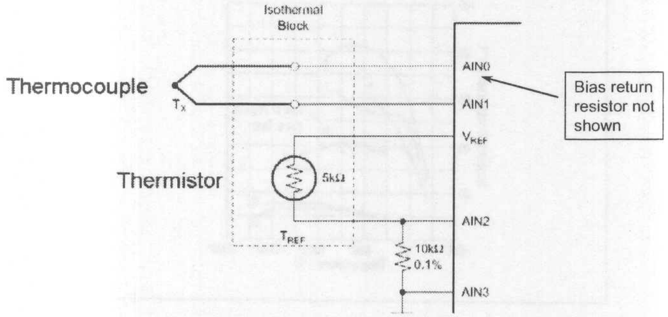
Thermocouples

Nonlinearity and compensation

Nonlinearity correction using An MSC1202 intelligent ADC

MSC1202 (See SBAS328)

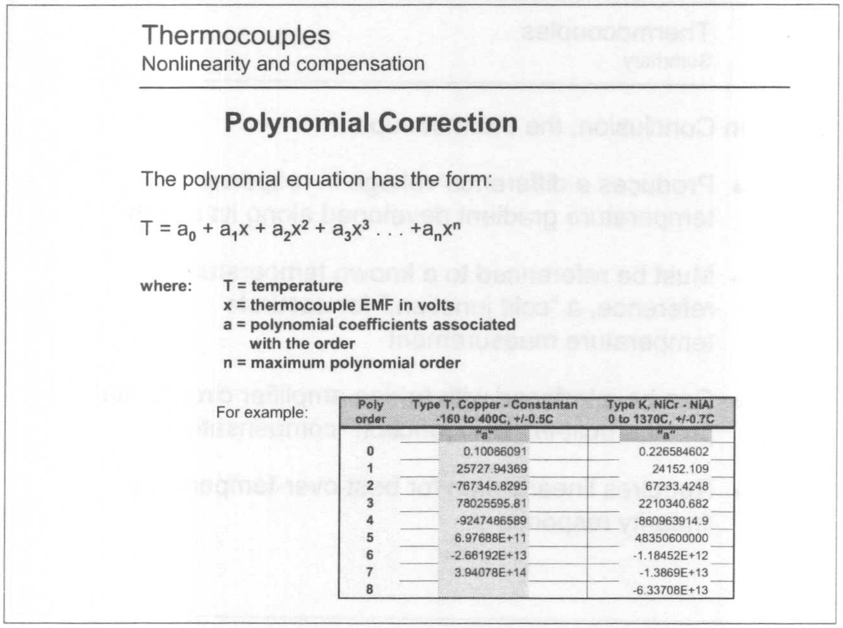
- **8051 CPU – 4kB flash memory**
- **16-bit ΔΣ ADS**
- **Current DAC**
- **8-differential or single inputs**



Bias return resistor not shown

This circuit does not use a linearization circuit for the thermistor, it simply uses a general-purpose equation to convert the resistance into a temperature. That temperature is then used to calculate the voltage for the thermocouple type which is used at that same temperature. This procedure calculates the voltage from 0°C to T_{REF}. The voltage is then added to the voltage measured from the thermocouple. The total voltage is then used to calculate the temperature at the end of the thermocouple.

See TI applications report SBAA134 for an extensive treatment of thermocouple temperature measurements with ΔΣ ADCs.



Thermocouples

Nonlinearity and compensation

Polynomial Correction

The polynomial equation has the form:

$$T = a_0 + a_1x + a_2x^2 + a_3x^3 \ldots + a_nx^n$$

where:
T = temperature
x = thermocouple EMF in volts
a = polynomial coefficients associated with the order
n = maximum polynomial order

For example:

Poly order	Type T, Copper - Constantan -160 to 400C, +/-0.5C	Type K, NiCr - NiAl 0 to 1370C, +/-0.7C
	"a"	"a"
0	0.10086091	0.226584602
1	25727.94369	24152.109
2	-767345.8295	67233.4248
3	78025595.81	2210340.682
4	-9247486589	-860963914.9
5	6.97688E+11	48350600000
6	-2.66192E+13	-1.18452E+12
7	3.94078E+14	-1.3869E+13
8		-6.33708E+13

Common thermocouples have been well characterized by the NIST and the applicable polynomial coefficients are available in the NIST's Thermocouple Tables (page Z-203). The polynomial order is established for a maximum error of ±1°C. The required order to achieve this will depend on the thermocouple type. If the application has a limited temperature range then a lower order polynomial correction will be sufficient.

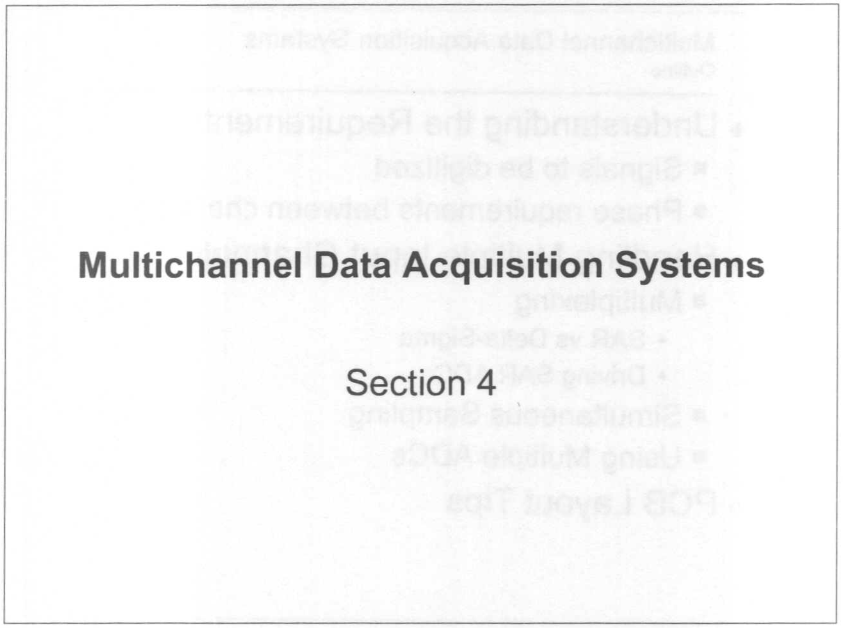
The mathematical expression shows how the polynomials are applied to the measured EMF (voltage). The tables lists as an example the coefficients for both a type-T and type-K thermocouple.

Thermocouples

Summary

In Conclusion, the thermocouple:

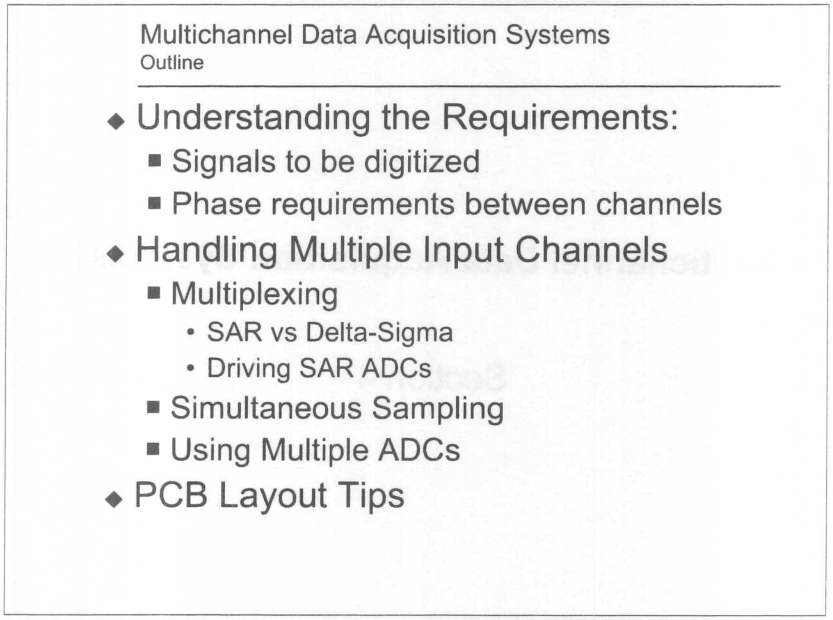
- Produces a difference voltage in response to a temperature gradient developed along its length
- Must be referenced to a known temperature reference, a “cold junction,” for accurate temperature measurement
- Can be interfaced with bridge amplifier circuits that provide built-in, “cold junction” compensation
- Requires linearization for best over-temperature linearity response



Multichannel Data Acquisition Systems

Section 4

This section will look at considerations for multichannel data acquisition systems.



Multichannel Data Acquisition Systems

Outline

- Understanding the Requirements:
  - Signals to be digitized
  - Phase requirements between channels
- Handling Multiple Input Channels
  - Multiplexing
    - SAR vs Delta-Sigma
    - Driving SAR ADCs
  - Simultaneous Sampling
  - Using Multiple ADCs
- PCB Layout Tips

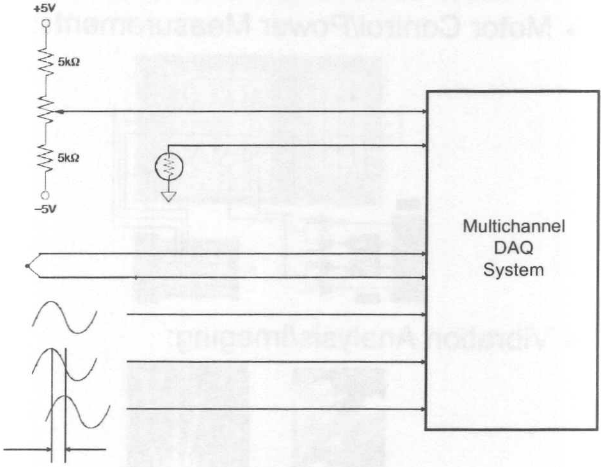
In this session, we'll be examining some of the issues present in designing a system with multiple channels. We must begin by knowing something about the signals we want to digitize, both in the frequency and the time domains. This will help us determine the approach that will best suit these requirements.

There are three main approaches: using a single converter and multiplexing each channel to it; using a similar approach but with individual sample/hold amplifiers on each channel to allow for the sampling instant of each channel to be simultaneous; and finally, using an individual ADC for each channel.

Along the way, we'll also look at the differences in multiplexing SAR and delta-sigma ADCs, issues with driving SAR ADCs, and some tips for preserving accuracy when creating a printed circuit board (PCB).

Multichannel Data Acquisition Systems

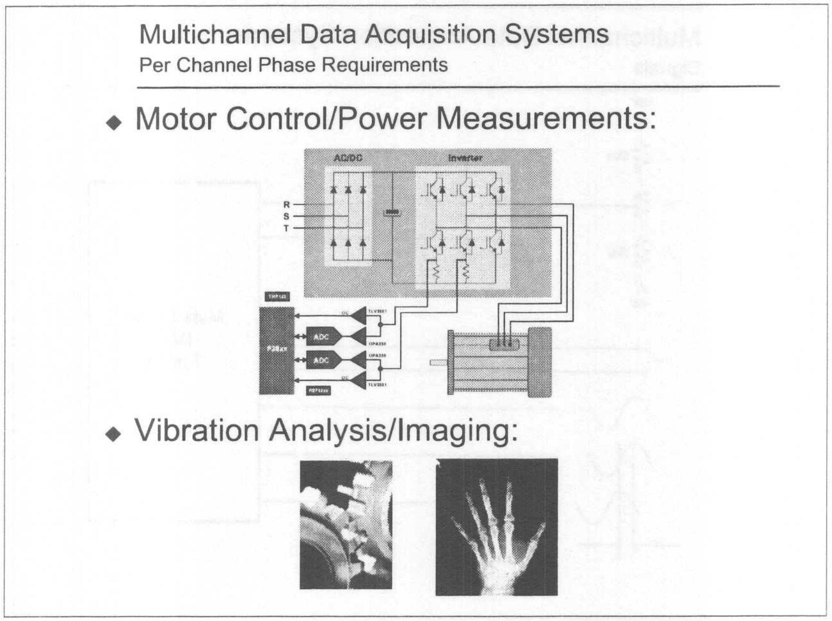
Signals



We must first know something about the signals we wish to digitize. Are they DC signals – perhaps from a number of thermistors or thermocouples? Or are they AC signals, perhaps the two microphones capturing a stereo signal? Or is a combination of both types of signals – perhaps the system needs to monitor power supply voltages while it measures power line frequency as well?

A multiplexed system is often the lowest cost, so it helps to look at that approach first. This means we must know the highest frequency of interest that is present in any one of the channels, as this may set our throughput requirements. But if each channel's needs are quite different, perhaps it makes more sense to use different converters that are more suited for each channel's requirements.

If the channels have some unique time relationship to each other, then simultaneous sampling may be required to preserve that phase information. This can be done with a multiplexed system and sample/holds, or perhaps it's easier with individual ADCs. We need to know more about these questions to determine what is the right approach.



Multichannel Data Acquisition Systems

Per Channel Phase Requirements

- Motor Control/Power Measurements:

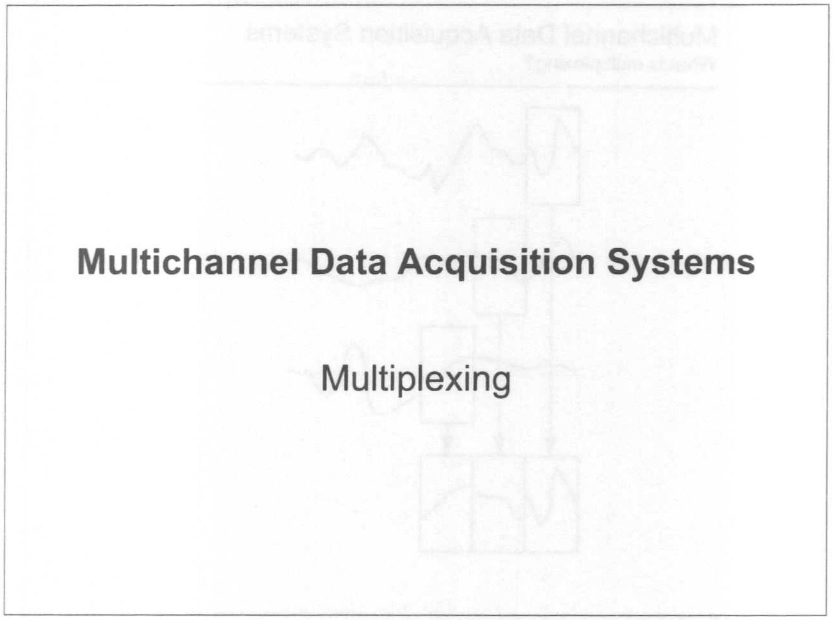
AC/DC
Inverter
R
S
T
TMP142
F28xx
ADC
ADC
TLV3501
OPA350
OPA350
TLV3501

- Vibration Analysis/Imaging:

One of the most strenuous requirements is preserving phase information between channels – particularly if the channels all must be sampled at the same time. For example, in a power measurement application, the voltage and current must be measured at precisely the same time in order to make an instantaneous power measurement and do it accurately.

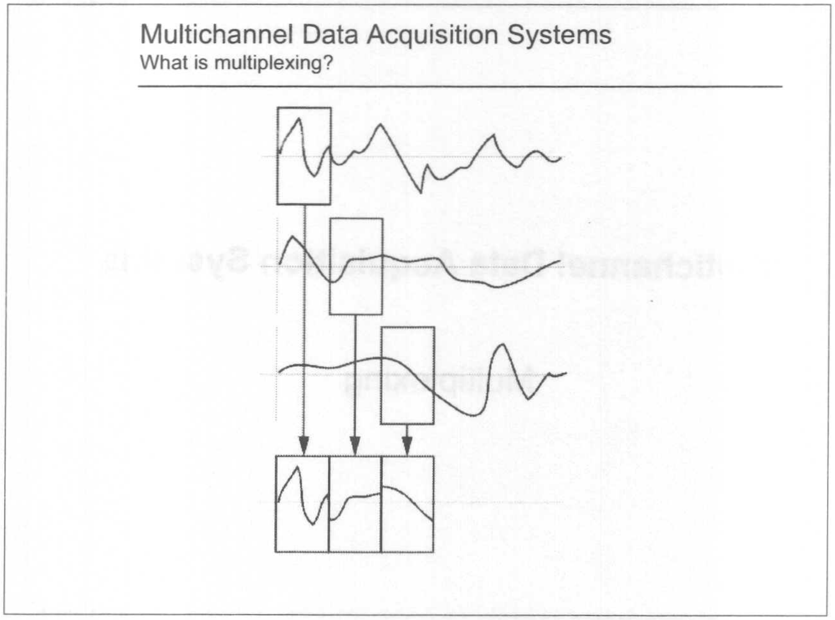
For some systems, such as vibration analysis and imaging, the phase information between channels is significant. This means that the delays between channels must be matched and the channels must likely be sampled at the same time. The challenge to the system designer is thus also one of matching the external filtering to within the phase margin allowed.

Which brings us to a point: what exactly does it mean to be "at the same time"? "Real-time" systems really aren't, but they operate at a frequency which makes them appear "real-time" - if you need to sample 50Hz signals simultaneously, does that require 0ns difference between channels, or can your system tolerate a few microseconds delay?



Multichannel Data Acquisition Systems

Multiplexing



Multichannel Data Acquisition Systems

What is multiplexing?

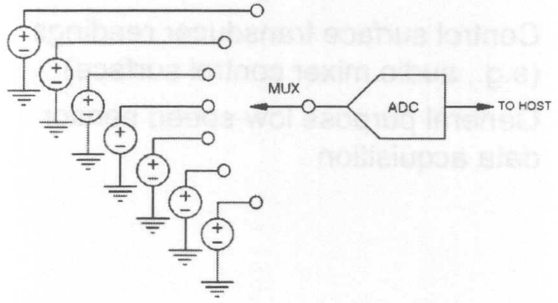
Multiplexing is the process of scanning through a number of input channels, and sampling each in rotation. In multiplexed systems, only one ADC is needed to acquire data from (potentially) many channels.

In a multiplexed system, multiple data streams are merged into a time-division multiplexed signal, which is then sampled.

Multichannel Data Acquisition Systems

Why multiplex?

- Fewer converters needed per channel – often only one needed
- Often lower power
- Often lower cost



Multiplexing allows one ADC to do the work of several in a multichannel system. Rather than dedicating an ADC to each channel, a single converter can read every channel in sequence. This can save power, since ADCs can use a significant amount of power while switches use very little, and it can save cost, since switches are much cheaper than ADCs.

Multichannel Data Acquisition Systems

Applications for multiplexing

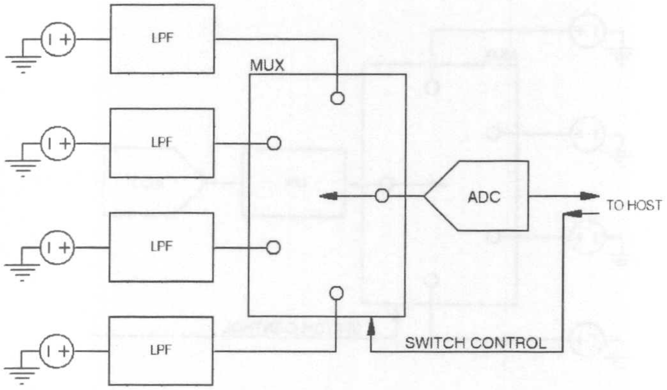
- Temperature monitoring
- Multi-point weigh scales
- Pressure monitoring
- Control surface transducer readings (e.g., audio mixer control surface)
- General purpose low-speed sensor data acquisition

Multiplexing cannot be used if channels must be sampled simultaneously. It also tends to limit the data rate, since the available throughput of a single converter is divided among many channels.

Note that some applications which seemingly require simultaneous sampling can still use multiplexing if the requirement is for constant phase between samples, for example, the phase relationship between channels need not be zero degrees, but can be some fixed number.

Multichannel Data Acquisition Systems

SAR-based multiplexed system



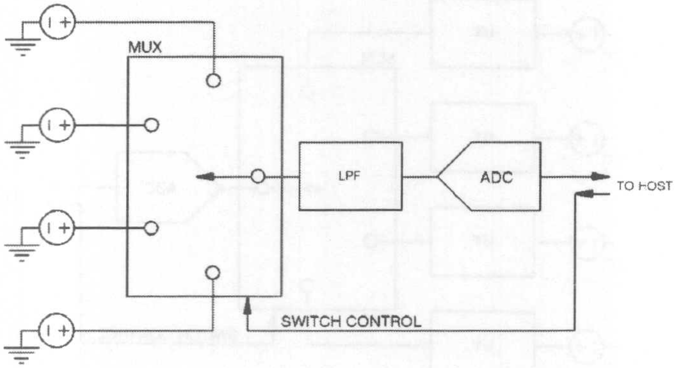
Multiplexing is typically done with SAR converters. The individual signals are anti-aliased before the multiplexer, and the SAR takes one sample from each time slot.

Aliasing at the converter is prevented by keeping the conversion time properly synchronised with the multiplexer switching.

As we will see, this method unfortunately cannot be used with delta-sigma converters.

Multichannel Data Acquisition Systems

Delta-sigma multiplexed system

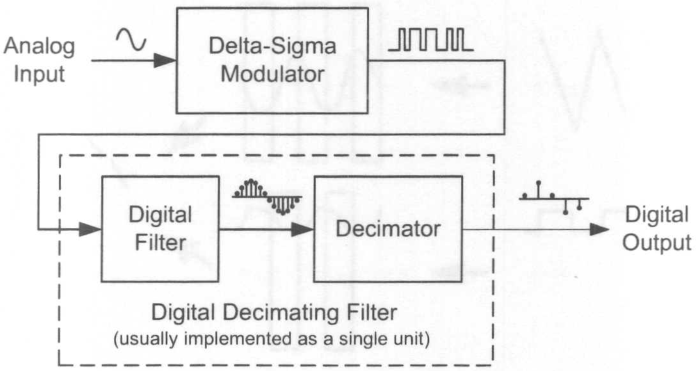


A multiplexed system using a delta-sigma converter typically has the form shown. The major change is that the anti-aliasing filter has been placed between the multiplexer and the ADC, and there are no longer anti-aliasing filters on each input channel.

There could be anti-aliasing filters on each input channel, but they would be redundant: the anti-aliasing filter in front of the ADC is an integral part of the delta-sigma.

Multichannel Data Acquisition Systems

Delta-sigma architecture



Analog Input

Delta-Sigma Modulator

Digital Filter

Decimator

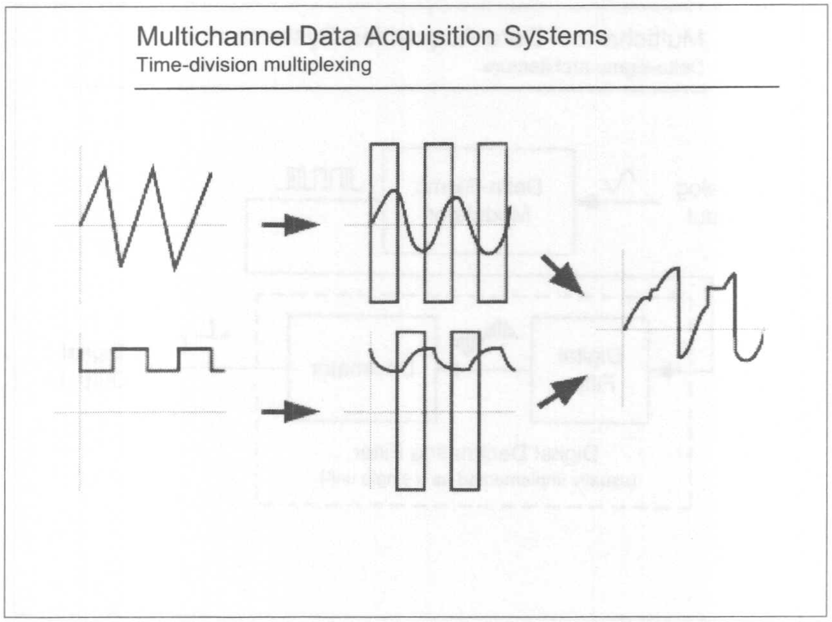
Digital Output

Digital Decimating Filter

(usually implemented as a single unit)

The low-pass filter is not optional in a (typical) delta-sigma converter. It is required to filter out the modulator's quantization noise. Its filtering action is also used to reject out-of-band frequencies that fall within the modulator's Nyquist bandwidth.

Because of this, in a delta-sigma system using a multiplexer, the multiplexer's switching action is always filtered, and we must consider the filter's settling time.



Multichannel Data Acquisition Systems

Time-division multiplexing

This diagram shows what a simple multiplexed signal might look like.

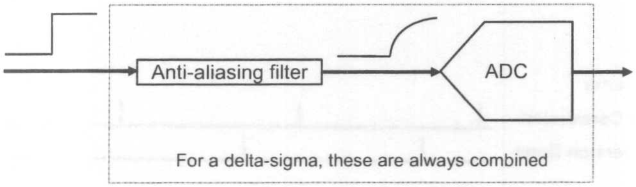
In this diagram, two signals are multiplexed. The signals have very different waveforms. A slice is taken of each, and the slices are concatenated alternately, as shown in the merged waveform. This waveform has high-frequency components – evidenced by the large and fast transitions – which are not present in the original signals; they are added by the multiplexing process.

A SAR converter can be configured to not sample these transitions, but the delta-sigma converter does not work this way. Its filter will react to the entire multiplexed waveform. This means that the fast transitions, which are like step functions, will be filtered, and the filter will have to “settle”. We will see this more clearly in the next slides.

Of course, a SAR converter can also be configured in the same way as a delta-sigma converter, by inserting the anti-aliasing filter between the multiplexer and the converter. If that is done, the issues become very similar.

Multichannel Data Acquisition Systems

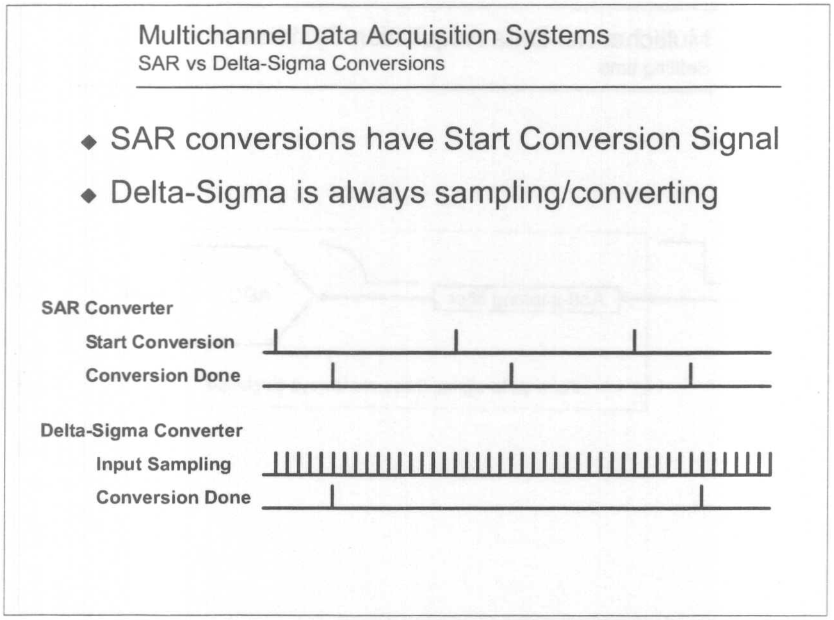
Settling time



Settling time is a property of any low-pass filter. A step input, when low-pass filtered, becomes a slow rise, possibly with a ringing phase; eventually this settles to the step input's final value. The length of time this takes depends on the filter, and is the major component of settling time in delta-sigma converters.

Delta-sigma converters incorporate a low-pass filter, so it must be taken into account when determining settling time. These digital filters are linear systems with many of the same parameters as continuous-time filters; they too have a step response. It may be very short, as with most industrial delta-sigmas, or it may be very long, as with most audio converters.

Note that for SARs, settling time means something slightly different; it refers to the settling that occurs during the acquisition time. This is different from settling time through the system as a whole, including the effect of the low-pass filter.



Multichannel Data Acquisition Systems

SAR vs Delta-Sigma Conversions

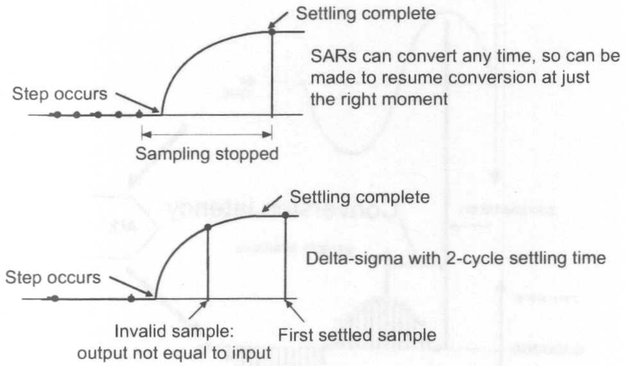
- SAR conversions have Start Conversion Signal
- Delta-Sigma is always sampling/converting

A SAR converter takes a “snapshot” of an input voltage and analyzes it to determine the corresponding digital code. A delta-sigma converter measures the input for a certain period of time and outputs a digital code corresponding to the average over that time. It is important to remember the way delta-sigma converters operate, particularly for designs incorporating multiplexing and synchronization.

It is very easy to synchronize delta-sigma converters together, so that they sample at the same time, but it’s more difficult to synchronize a delta-sigma converter to an external event.

Multichannel Data Acquisition Systems

Settling time: SAR vs. Delta-Sigma



Settling complete

SARs can convert any time, so can be made to resume conversion at just the right moment

Step occurs

Sampling stopped

Settling complete

Delta-sigma with 2-cycle settling time

Step occurs

Invalid sample: output not equal to input

First settled sample

SARs have an advantage over delta-sigmas in one respect; their sample times can be made arbitrary, within device limits. Sampling can be made to cease until the precise moment when the filter has recovered from a step input.

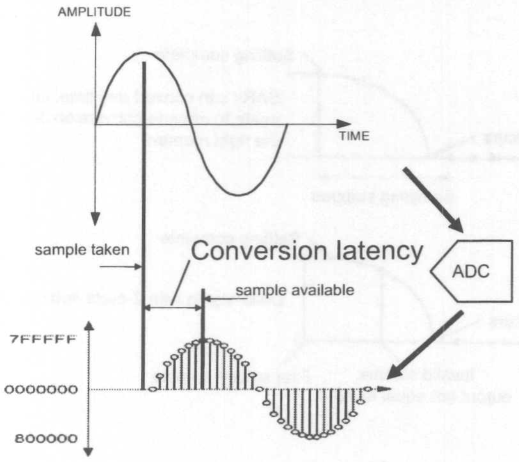
Delta-sigmas are generally made so that one conversion period follows another immediately, so as to maximize throughput. You can start a conversion early by resetting the converter, so starting it over, but this causes the digital filter to refill and resettle. Some time is nearly always wasted, unless the settling time is exactly matched to the data rate and a full-scale step has occurred at the right moment.

For a delta-sigma, settling time can never be less than the conversion latency, but often is more, and is always measured in conversion cycles. For a SAR, settling time depends very little on the converter itself, and very much on the antialiasing filter's step response time – if there is one. SARs can be operated with no filter at all, if desired, and this is occasionally appropriate.

In multiplexed systems, this settling time is very important; there could be a maximum-scale step at every multiplexer transition. Therefore, the output of the delta-sigma must be considered invalid until the filter settling time has passed.

Multichannel Data Acquisition Systems

Latency



Latency and settling time are very commonly confused, but they are not at all the same thing. In fact, they have nothing to do with each other. Converter latency in multiplexed systems is far less important than settling time, being largely irrelevant to the process of multiplexing.

Conversion latency for an ADC is the amount of time that passes between the taking of a sample and the time that the sample is ready for retrieval.

For SAR ADCs, latency is typically very short, being the amount of time for the successive approximation process to complete. Data is typically available immediately afterward.

For delta-sigma ADCs, latency is a bit harder to define, since delta-sigma ADCs do not output a code corresponding to a single point in time. The code they output is the result of filtering or averaging the input during an interval of time; the interval is equal to the sample period. (Actually, this is also true for SARs, but the interval is extremely small compared to the sample period.)

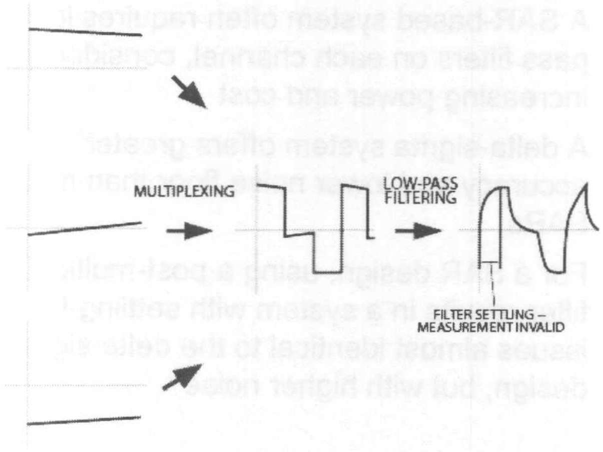
For this reason, we measure latency for a delta-sigma ADC by starting at the beginning of a sample period, and measuring to the time that data can be retrieved. It may also be practical to include in this the time needed to retrieve the data, since delta-sigma ADCs nearly always have serial interfaces.

Delta-sigma ADCs have an additional “disadvantage” compared to SARs; they always include a filter. Some latency may be incurred in the filter. For audio converters, this additional latency can be very significant, even up to several tens of sample periods. For low-speed industrial converters with sinc filters, it sometimes amounts to only a few modulator cycles.

The term “group delay” is often used in place of latency. For SARs, which generally have a continuous-time filter at the front-end, group delay can be a more useful measurement, since the latency may be frequency-dependent. For delta-sigma ADCs, filters with constant group delay are almost always used, so there is no difference with latency.

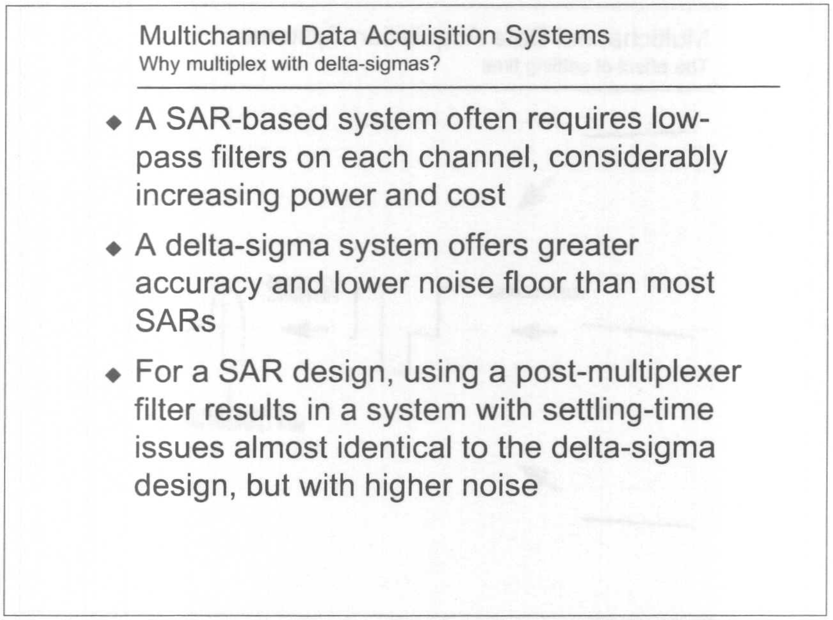
Multichannel Data Acquisition Systems

The effect of settling time



The effect of settling time on a multiplexed system is shown in the diagram.

Here we show three near-DC signals multiplexed into a single stream, which is then low-pass filtered. The multiplexer switches are filtered along with the individual signals, and this produces “settling”. After each channel switch, measurements must be withheld until this settling time passes. This reduces per-channel throughput, but each channel is effectively anti-aliased.



Multichannel Data Acquisition Systems

Why multiplex with delta-sigmas?

- A SAR-based system often requires low-pass filters on each channel, considerably increasing power and cost
- A delta-sigma system offers greater accuracy and lower noise floor than most SARs
- For a SAR design, using a post-multiplexer filter results in a system with settling-time issues almost identical to the delta-sigma design, but with higher noise

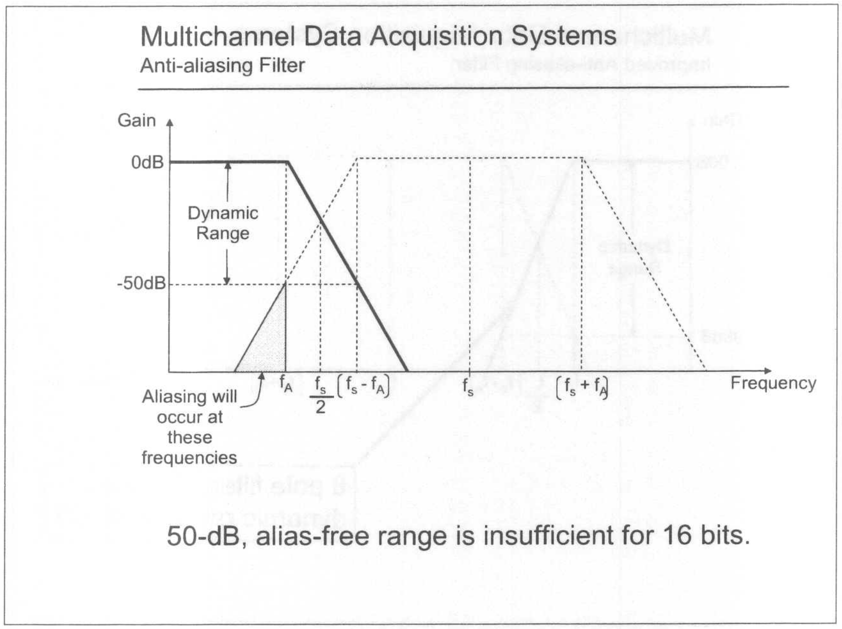
Many people get discouraged when they consider the settling penalty incurred for some delta-sigma filters. Despite this penalty, there are very compelling reasons to use a delta-sigma converter in a multiplexed design.

The primary reason is accuracy and noise; few SAR converters can even approach the noise floor possible with a delta-sigma converter. In systems where this is very important, the reduced throughput of the delta-sigma ceases to be an issue.

Furthermore, a SAR system often requires low-pass filtering on each channel, and since a SAR has no anti-aliasing at all, these filters may need to be complex multi-pole active designs. These can reduce accuracy and consume considerable power.

If this problem is circumvented by using only a single complex filter after the multiplexer, you obtain a design very similar to the delta-sigma configuration; but the noise and accuracy issues remain.

The cycles lost to settling time aren’t the critical issue in most designs; the real issue is overall throughput for a given level of accuracy, and on these counts the delta-sigma design often wins by a considerable margin.



Multichannel Data Acquisition Systems

Anti-aliasing Filter

50-dB, alias-free range is insufficient for 16 bits.

An anti-aliasing filter, with low-pass characteristics, may be placed in the signal chain — in front of the ADC. The purpose of the filter is to attenuate any high-frequency components in the signal and to ensure that the samples taken by the ADC satisfy Nyquist.

Consider the example shown here:

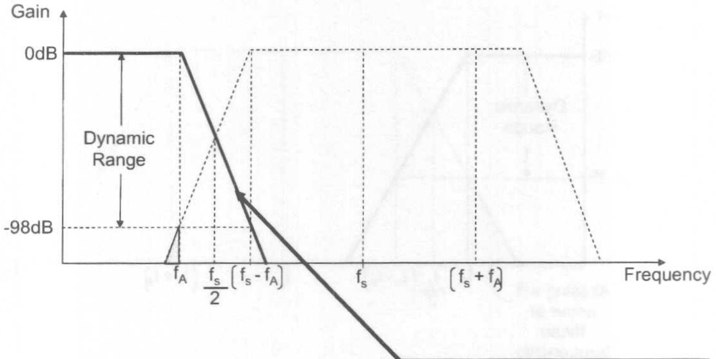
Suppose we have an ADC with a spectrum shown above and a maximum full-scale frequency content of 40 kHz. Also, at 70 kHz, the signal is attenuated by 50 dB (f_S-f_A).

We can see from inspection that at the frequency f_A, the alias component, limits the system's dynamic range to 50 dB.

The system requires an additional 48 dB of attenuation at 70 kHz to achieve the 98 dB of attenuation required for 16-bit performance.

Multichannel Data Acquisition Systems

Improved Anti-aliasing Filter



8 pole filter increases dynamic range to 98 dB

If the application requires additional dynamic range, you will have to consider an anti-aliasing filter to provide more attenuation at 70 kHz.

For a 16-bit system, remembering that SNR = 6.02n + 1.76 dB, we arrive at an SNR of around 98 dB.

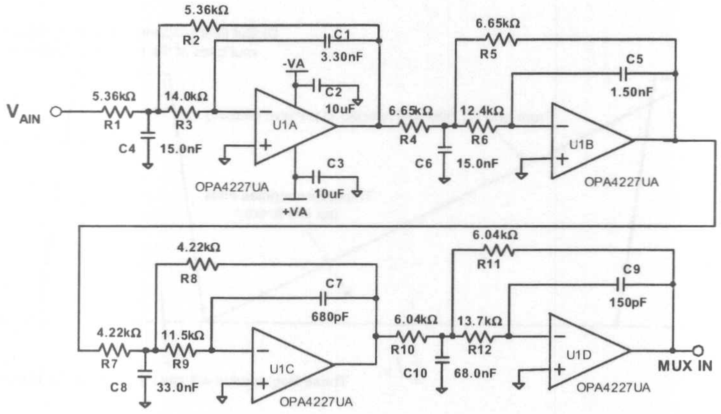
So, if you really want true 16-bit performance at 40 kHz, you need to add a filter that gives 48 dB of attenuation at 70 kHz.

From 40 kHz to 70 kHz almost doubles the frequency—this is an increase of one octave. Each filter pole will give 6 dB per octave. You will need an 8-pole filter to limit the input frequency of the signal and ensure the full dynamic range at the frequency of interest.

In applications that would require an anti-aliasing filter, the signal's spectral characteristics and the system's dynamic-range requirements must be completely understood before beginning.

Multichannel Data Acquisition Systems

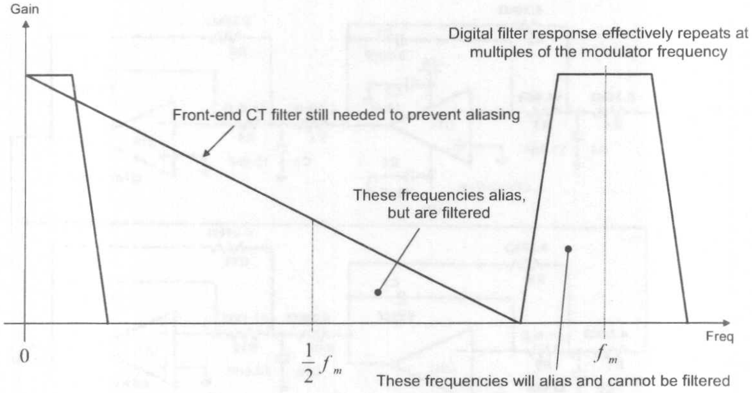
Anti-aliasing Filter Circuit



An 8th order filter circuit is shown here. A quad op amp makes this complex looking circuit relatively small. Still, the cost of the circuit is approximately $5 – as much or perhaps more than the converter itself!

Multichannel Data Acquisition Systems

Delta-sigma digital filters



Digital filter response effectively repeats at multiples of the modulator frequency

Front-end CT filter still needed to prevent aliasing

These frequencies alias, but are filtered

These frequencies will alias and cannot be filtered

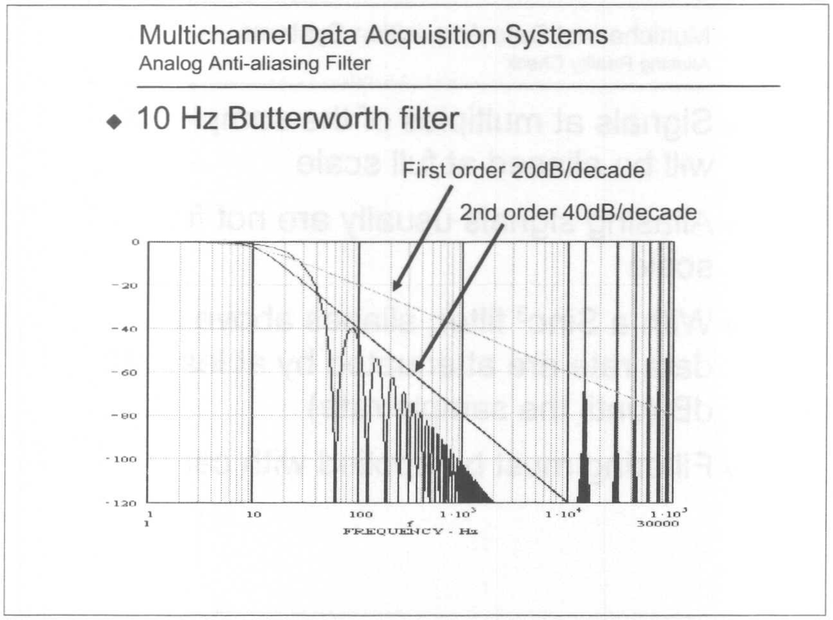
It is widely believed that delta-sigma converters do not require anti-aliasing filtering on the front end. While this may happen to work sometimes, in general it is not the case.

The problem is that a digital filter operates on a discrete-time signal, which contains in itself no information about time. A discrete-time signal sampled at a given rate contains all frequencies aliased into the Nyquist bandwith. What this means in the continuous-time domain is shown in the figure.

To be sure, in many circumstances the modulator frequency is at a point where there is little information at the Nyquist rate and above; but noise will always be present, and this aliases into the bandwidth of interest just as any other energy will, raising the final noise floor.

Worse still, if there is energy present around the modulator frequency, it will alias down: the digital filter is powerless to remove it, since from its point of view, these frequencies are in the baseband already. It is impossible for it to tell the difference.

For this reason, a continuous-time filter is often needed in front of a delta-sigma. Its stopband attenuation should at least begin at the point shown above. This is especially true for high-order delta-sigmas where the noise floor of the converter itself is very low. Fortunately, this filter often need not be very strong, especially if the digital filter has a very low cutoff frequency. This is why even a simple RC filter can sometimes suffice for a delta-sigma.



Multichannel Data Acquisition Systems

Analog Anti-aliasing Filter

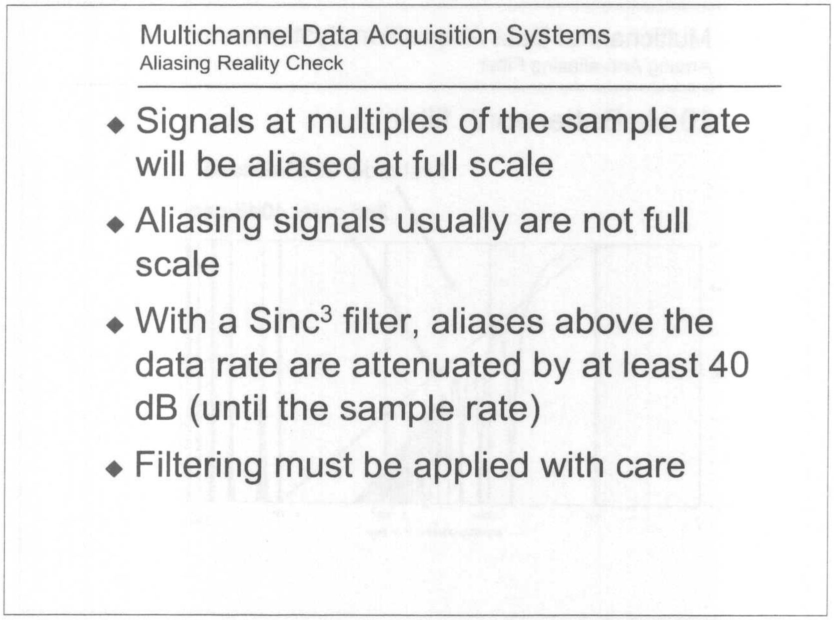
- 10 Hz Butterworth filter

First order 20dB/decade

2nd order 40dB/decade

A single-pole RC anti-aliasing filter may work in some delta-sigma applications, but even here a stronger filter is sometimes needed, as in the above example.

You can see that by adding a single pole filter, signals at the modulator frequency may not be attenuated sufficiently. Using a two-pole filter may be necessary.



Multichannel Data Acquisition Systems

Aliasing Reality Check

- Signals at multiples of the sample rate will be aliased at full scale
- Aliasing signals usually are not full scale
- With a Sinc^3 filter, aliases above the data rate are attenuated by at least 40 dB (until the sample rate)
- Filtering must be applied with care

Of course, any time you consider an antialising filter, you may not have to assume the worst case – signals that may alias are likely not at full scale, but may be attenuated already by the sensor or system response. This is why it helps to know something about the nature of the signal you are attempting to digitize; it helps you know how much filtering is really required.

Multichannel Data Acquisition Systems

Calculating throughput

- Throughput: number of valid samples per second on all channels
  = converter data rate / filter settling time in cycles
- Per-channel data rate: number of valid samples per second on each channel individually
  = Throughput divided by number of channels to scan per cycle

For a multiplexed delta-sigma system, the throughput is the number of valid samples per second that can be obtained when a filter resynchronization is performed after every multiplexer switch. It is calculated by dividing the converter's data rate by the number of conversion cycles occupied by the filter's settling period. Note that this number does not depend on the number of channels to be scanned per cycle.

The actual sampling rate per channel depends on how many channels are to be scanned, and is obtained by dividing the throughput by the number of channels to be scanned.

Multichannel Data Acquisition Systems
Choosing a delta-sigma for multiplexing

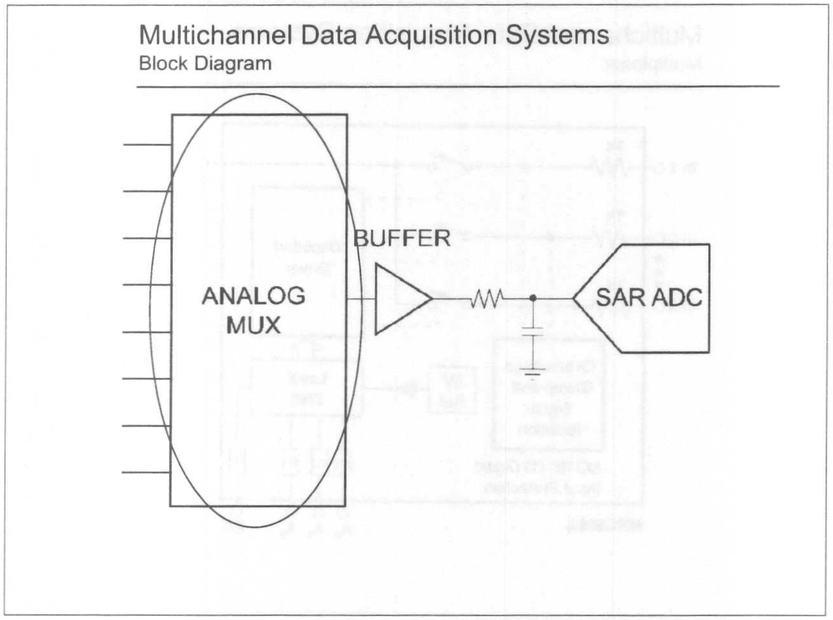
- Important specifications:
  - Settling time: lower is better
- Unimportant specifications:
  - Latency / group delay
    - Watch out – occasionally settling time is called latency in datasheets!
- Important characteristics:
  - Ease of configuration & channel switching

We will now consider a few delta-sigma devices and their suitability for use in multiplexed systems.

The important specification is filter settling time. If it is low, the part will be efficient in a multiplexed system.

Latency and/or group delay are **not** important for multiplexing. A part with a high latency may be good for multiplexing (although admittedly this is unusual, since high latency filters typically also have a long settling time).

A secondary consideration, but possibly of importance, is the ease with which a device can be used in a multiplexed system.



Multichannel Data Acquisition Systems

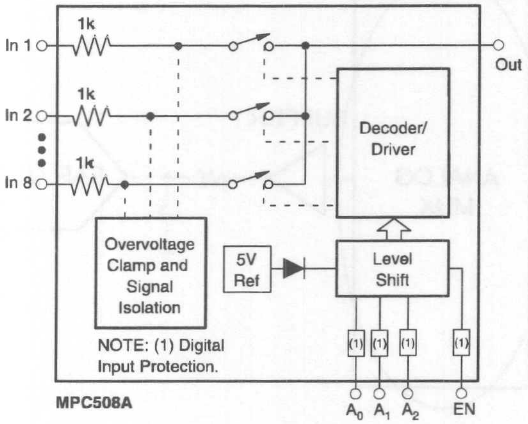
Block Diagram

A quick block diagram helps to frame the multiplexed system design. Each channel will require its own signal conditioning and an antialiasing filter, and then go into the multiplexer. The multiplexed signal will then be routed to the A/D converter through some type of buffer. We'll route the resulting digital data to the system processor.

Because the system is general purpose, we really don't know what kinds of signals or applications the measurement system may be required to handle. If the system were to be used in a high speed control loop, the time from measuring a channel to getting a precision digital representation of the signal should be minimized. Why are we concerned with this? It helps us to choose the converter type to use, as in the previous discussion of SAR versus delta-sigma converters.

Multichannel Data Acquisition Systems

Multiplexer

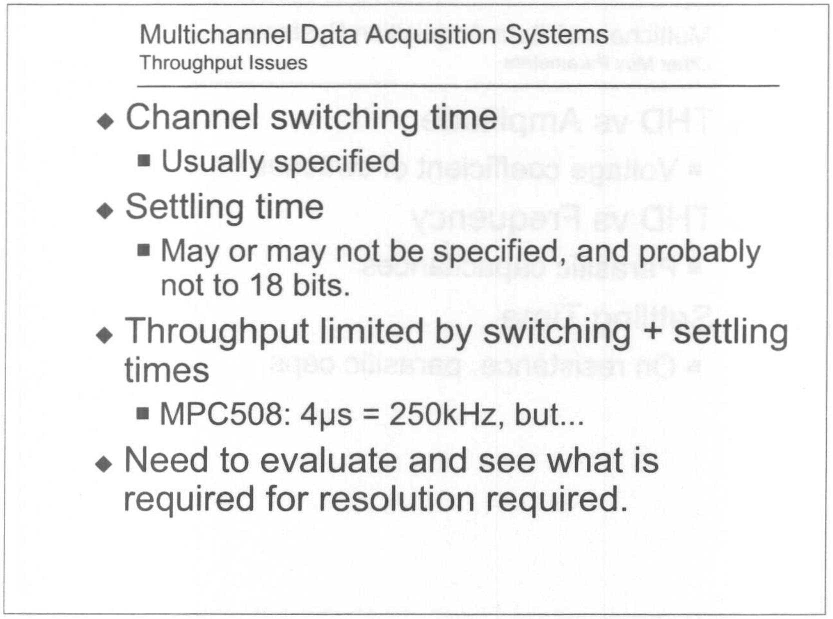


NOTE: (1) Digital Input Protection.

A multiplexer is simply a set of analog switches which have one common terminal, and can be easily addressed and turned on under digital control.

The first important parameter for a multiplexer is obviously the number of channels that it has, and whether or not it handles single-ended or differential inputs. In the system shown previously, the inputs are all single-ended, since our signal conditioning and filtering have provided us with single-ended signals. So a simple single-ended 8–channel multiplexer is all that is required.

The analog switches in a multiplexer may have significant on-resistance – in the one shown above, it is 1kΩ. Because of this, the output of the multiplexer is usually buffered with an op amp.



Multichannel Data Acquisition Systems

Throughput Issues

- Channel switching time
  - Usually specified
- Settling time
  - May or may not be specified, and probably not to 18 bits.
- Throughput limited by switching + settling times
  - MPC508: 4μs = 250kHz, but...
- Need to evaluate and see what is required for resolution required.

Since the multiplexer is the device responsible for switching from one channel to the next, another important parameter is the time it takes for the mux to switch from channel to channel. Once switched to a new channel, how long does it take for the output to settle?

The first parameter is often shown on a mux datasheet. Looking at the MPC508 which was shown on the previous slide, it takes 0.5μs to change from one channel to the next. Its settling time is only listed to 0.01% - and this takes 3.5μs! So switching from one channel to another would take at least 4μs – and that is only to 12 bits! Some experimentation is needed to determine the settling time required for a system with higher resolution.

Multichannel Data Acquisition Systems

Other Mux Parameters

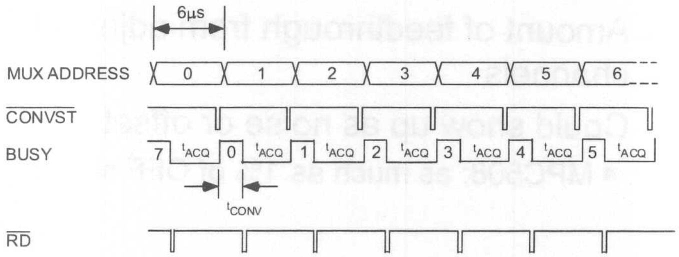
- THD vs Amplitude
  - Voltage coefficient of switches
- THD vs Frequency
  - Parasitic capacitances
- Settling Time
  - On resistance, parasitic caps

Some other parameters to consider on multiplexers are the dynamic performance when switched to a particular channel – especially total harmonic distortion (THD). Multiplexer switches have a significant number of parasitic capacitances which can affect the signal from a THD vs frequency and amplitude perspective.

From a more “industrial” or DC standpoint, the settling time of the multiplexer is significant, as this directly affects the throughput.

Multichannel Data Acquisition Systems

System Timing – First Look

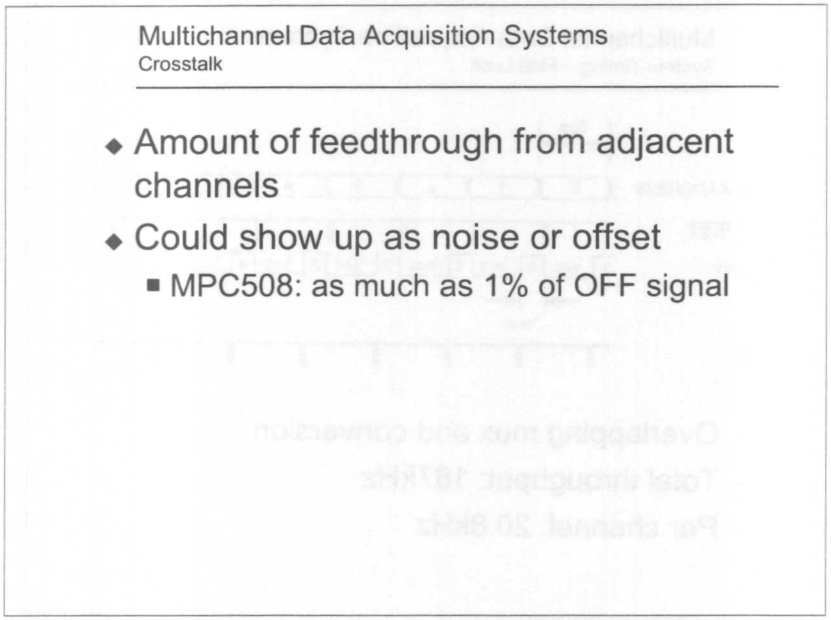


- Overlapping mux and conversion
- Total throughput: 167kHz
- Per channel: 20.8kHz

We need to determine just what throughput our system will have, as this will affect decisions we make about other parts of the system.

Because we control the time at which the multiplexer switches, and the time the converter starts its conversion, we can take into account the overall settling of the multiplexer and its buffer op amp, if used. Shown here is the timing diagram of how we command the multiplexer to switch and start conversions. Assuming 6μs is required for full settling. The key here is to switch the multiplexer, wait 6μs, then start a conversion. Once the conversion has started, the A/D converter has the input held on its internal sampling capacitor, so we are free to change the multiplexer channel, allowing it to begin settling while the converter is converting the previous channel's data that it has acquired. Once that conversion is complete, we read it as quickly as possible, so that the reading of the data is not occurring during the critical acquisition time of the ADC, which could potentially cause noise and corrupt the acquired signal.

From this, we can see the overall throughput that is attainable by our system; 167kHz sample rate, which equates to a 20.8kHz rate per channel if we were scanning through all eight channels. However, if we wanted to switch to one channel and stay there, then we wouldn't have to allow for mux settling, and could potentially sample at a higher rate.

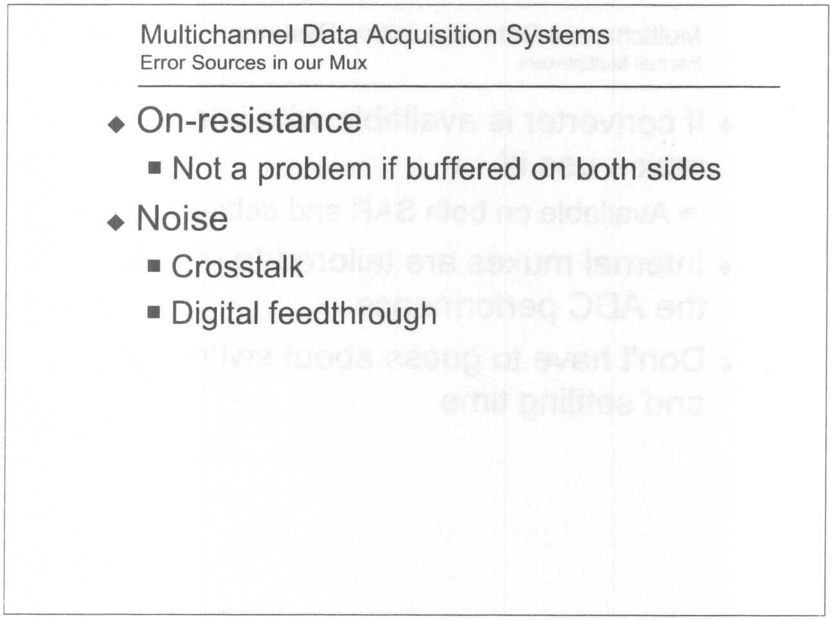


Multichannel Data Acquisition Systems
Crosstalk

- Amount of feedthrough from adjacent channels
- Could show up as noise or offset
  - MPC508: as much as 1% of OFF signal

A further specification that is often shown on multiplexer datasheets is crosstalk. This is the amount of signal that may feed through from 'off 'channels when one channel is selected. This would commonly show up as noise in the channel of interest, but may also manifest itself as an offset, depending upon the signals used and the particular multiplexer in use.

Again, this may or may not be specified. If it is, it may be in a curve plotted against signal frequency – higher frequencies will tend to feed through more as the mechanism of crosstalk is usually parasitic capacitances associated with the analog switches.



Multichannel Data Acquisition Systems

Error Sources in our Mux

- On-resistance
  - Not a problem if buffered on both sides
- Noise
  - Crosstalk
  - Digital feedthrough

Since multiplexers are simply switches, and not so much an "active" component, they do not specify offset or gain errors. The main error source with multiplexers is their on-resistance, which is easily handled by driving them with an op amp and buffering the output with an op amp. In most systems, the antialiasing filter op amp drives the mux input, and the ADC buffer, buffers the output of the multiplexer, so this is not a concern.

Noise, however, can be a problem with multiplexers as the crosstalk and feedthrough of digital signals into the analog path can contribute noise.

Multichannel Data Acquisition Systems

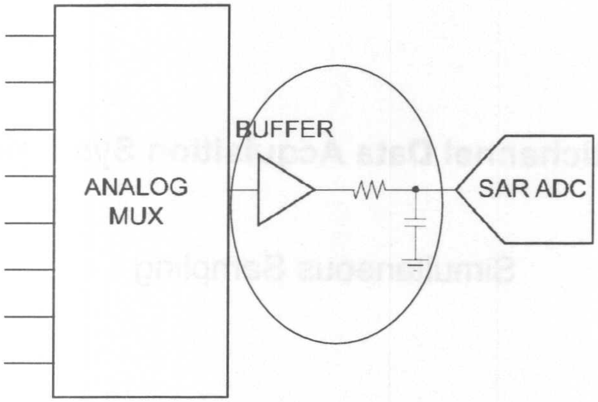
Internal Multiplexers

- If converter is available with internal mux – use it!
  - Available on both SAR and delta-sigma
- Internal muxes are tailored to match the ADC performance
- Don't have to guess about switching and settling time

If an ADC that meets your system needs is available with an internal multiplexer, this is by far the best solution. Not only does it save space on the board, but the multiplexers internal to the ADC are designed and optimized to work well with that particular ADC. This eliminates a lot of the guesswork that comes with using an external mux.

Multichannel Data Acquisition Systems

Block Diagram



Now that we've looked at the multiplexer, let's take a closer look at the buffer requirements and input circuitry needed for our ADC. SAR ADCs in particular require the RC circuit shown here. Choosing the values for the R and C as well as choosing the right buffer amplifier are important in preserving the accuracy of the system.

Multichannel Data Acquisition Systems

Simultaneous Sampling

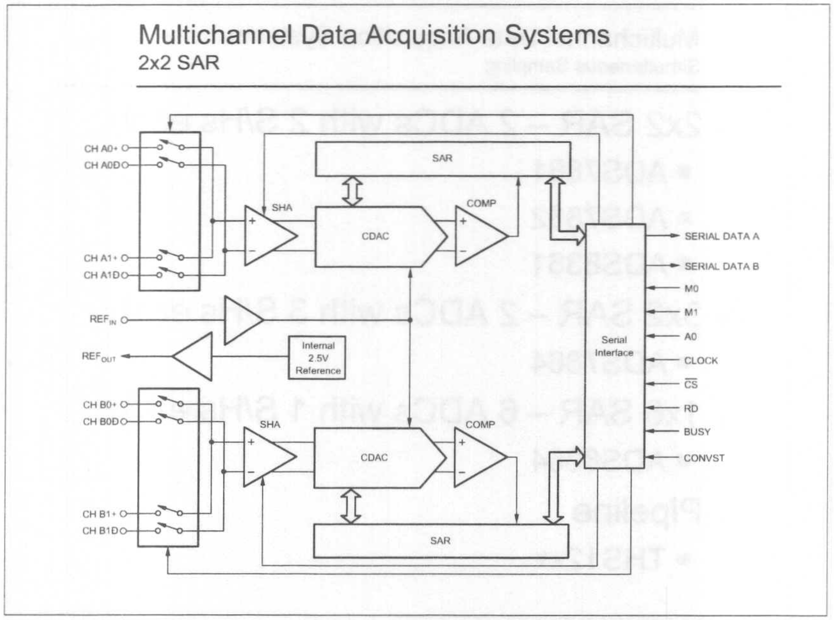
Multichannel Data Acquisition Systems

Simultaneous Sampling

- 2x2 SAR – 2 ADCs with 2 S/Hs each
  - ADS7861
  - ADS7862
  - ADS8361
- 3x2 SAR – 2 ADCs with 3 S/Hs each
  - ADS7864
- 1x6 SAR – 6 ADCs with 1 S/Hs each
  - ADS8364
- Pipeline
  - THS12xx

As noted earlier, simultaneous sampling involves using individual sample/hold (S/H) amplifiers for each channel, then into a mux or individual converter. As it is difficult to purchase sample/hold amplifiers today, and is virtually impossible to get them with 16-bit accuracy, this scheme generally requires converters specially made to do this function.

We offer several types, and refer to them in an MxN way, where M is the number of sample/holds, and N is the number of converters.



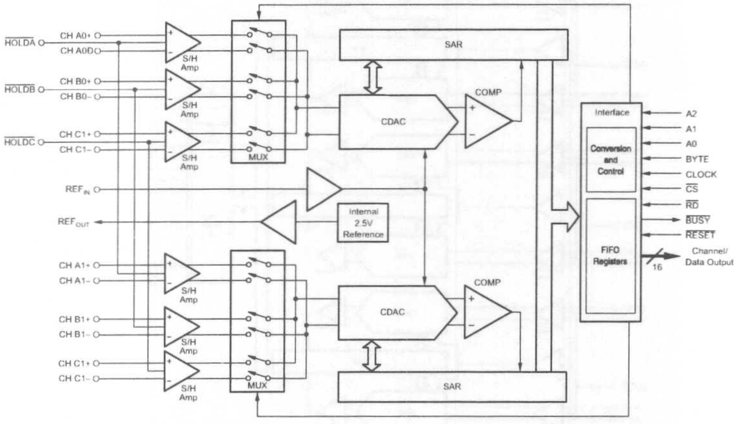
Multichannel Data Acquisition Systems

2x2 SAR

This 2x2 system uses two ADCs, each with a 2-channel mux before the sample/hold. Thus, two channels may be sampled at the same time, then the other two channels selected and sampled at the same time.

Multichannel Data Acquisition Systems

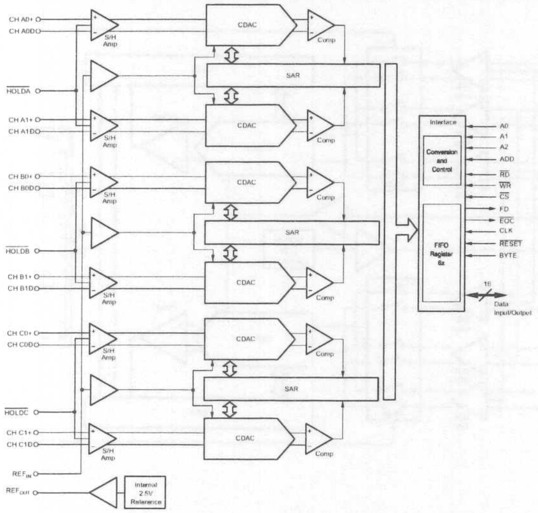
3x2 SAR



This 3x2 system puts the sample/hold ahead of the mux, so that all six channels can be sampled at the same time, then each converter can convert three of the channels one at a time.

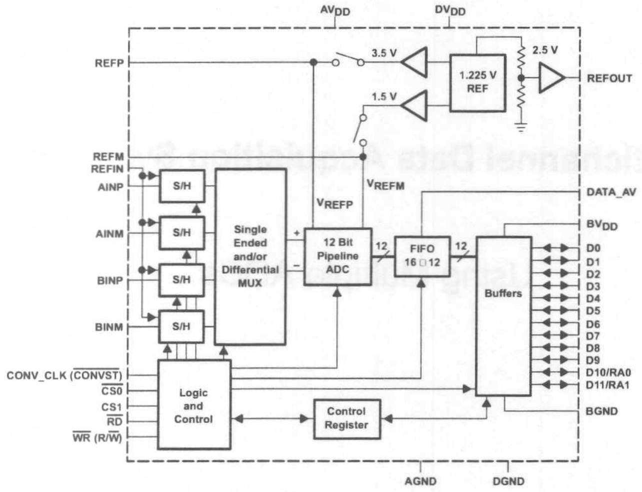
Multichannel Data Acquisition Systems

1x6 SAR



The 1x6 SAR is basically a converter-per-channel system, but all included in one IC. Each channel has its own converter and sample/hold, so no analog multiplexing is required.

Multichannel Data Acquisition Systems
Pipeline



This is the THS1206, which is what we would call a 4x1 converter – 4 S/Hs multiplexed to one converter.

Multichannel Data Acquisition Systems

Using Multiple ADCs

Multichannel Data Acquisition Systems

Multiplexed vs Separate Converters

- If one channel has very different needs than all the others:
    - Consider using a separate ADC for that channel, and mux the others
- Multiplexing implies time delay between channels – is that OK? Or can you use simultaneous sampling?
- Individual ADCs/channel is great, but can the digital side handle it?
    - And can you match the analog, if necessary?

If your system consists of 7 thermocouples and 1 ultrasonic sensor, perhaps it doesn't make sense to try and multiplex all of those signals. Would it make more sense to use a high-speed ADC for the ultrasound path, and a multiplexed, slower ADC for the relatively slow temperature measurements? These days, the costs of ADCs have come down low enough that such a tradeoff is possible, and often results in higher performance. If you muxed those signals, you may have trouble keeping crosstalk from the ultrasonic channel from affecting your temperature measurements.

As we've seen, multiplexing means we'll only look at one channel at a time, so that there is a delay between the results of one channel vs. another. Within the time requirements of the system, is that OK? Or is simultaneous sampling required – and is it just sampling that's needed, or do you need the data all at the same time?

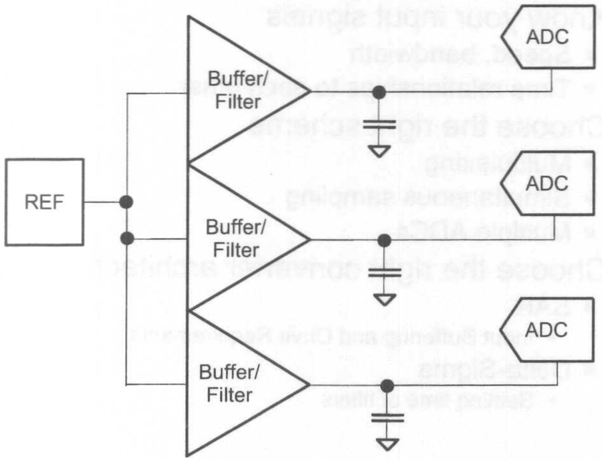
If so, you may need to consider using separate ADCs per channel. While this sells more ADCs for TI, there are some practical considerations that may advise against this. If you have all the converters convert at the same time, and your processor needs to read them all at the same time, does the processor have the resources (parallel/serial data ports) to do it? And if there's many channels and their time alignment must match, how confident are you that the signal conditioning matches, from channel-to-channel? How about over time/temperature? One nice thing about individual ADCs is you could calibrate them all separately to compensate for offsets and drifts of each channel – but this would be a very complex system.

Multichannel Data Acquisition Systems

Other Considerations

Multichannel Data Acquisition Systems

Reference Distribution



Reference distribution in multiple ADC systems is an interesting problem Certainly you wouldn't want to have multiple references if the results of the converters are all going to be compared and used in the same data set. Therefore, the system should have one central reference and the buffer/filter circuit. If only one buffer is used, make sure to have large bulk capacitors near the reference pin of each ADC.

Note that several buffers may be needed, as shown above, depending upon the system. Consider the reference loading of each converter, and buffer accordingly.

Multichannel Data Acquisition Systems

Summary

- Know your input signals
  - Speed, bandwidth
  - Time relationships to each other
- Choose the right scheme
  - Multiplexing
  - Simultaneous sampling
  - Multiple ADCs
- Choose the right converter architecture
  - SAR
    - Input Buffering and Drive Requirements
  - Delta-Sigma
    - Settling time of filters

Bridge Measurement Systems

Section 5

Bridge Measurement Systems
Outline

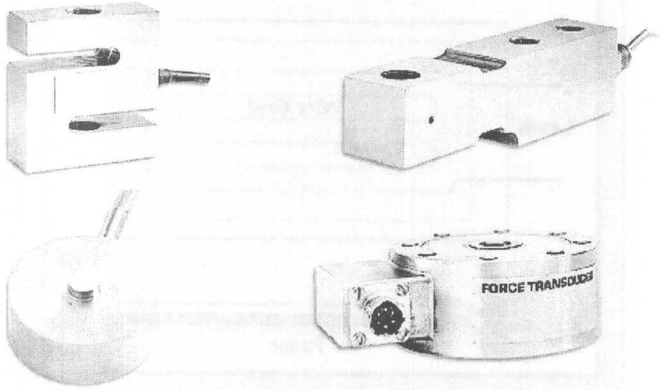
- Introduction to Bridge Sensors
- Circuits for Bridge Sensors
- A real design: the ADS1232REF
- The ADS1232REF Firmware

This presentation gives an overview of data acquisition for bridge sensors.

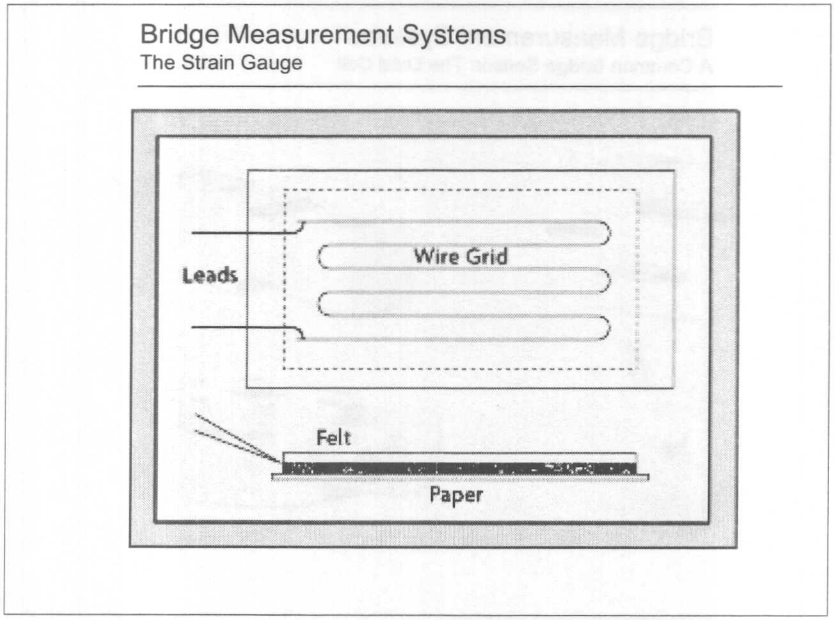
We begin with a discussion of the bridge sensor, and introduce some theory. We then look at actual hookups of bridge sensors to ADCs. The ADS1232REF is introduced and discussed as an example of a real design. Finally, we briefly discuss the ADS1232REF's firmware.

Bridge Measurement Systems

A Common Bridge Sensor: The Load Cell



A load cell is one of the most common bridge sensor types. It can measure most kinds of force, but load cells in weigh scales almost always measure compression.



Bridge Measurement Systems

The Strain Gauge

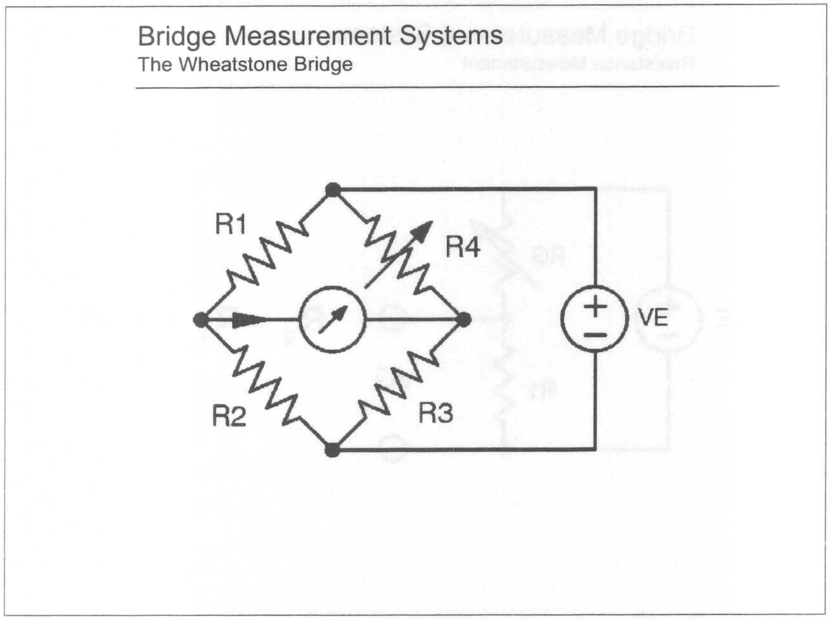
The strain gauge is the sensitive element in load cells.

Any common piece of wire will change resistance slightly when placed under tension. Wire arranged to maximise this effect forms a strain gauge. Strain gauges are typically etched foil patterns glued to a bar of metal or other tensionable material.

In a load cell, the strain gauge is placed so that when the cell is loaded, the gauge is strained, changing its resistance.

The resistance change per unit of strain is very small, and requires sensitive circuitry to measure it accurately.

(Image from http://www.omega.com/literature/transactions/volume3/strain.html)



Bridge Measurement Systems

The Wheatstone Bridge

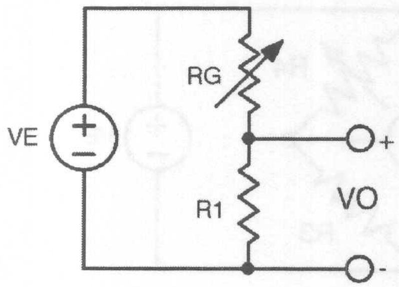
The Wheatstone bridge is a familiar circuit for measuring resistance very accurately. Invented in 1833 by Hunter Christie, the circuit was later studied by Charles Wheatstone, whose name became attached to the circuit thanks to his extensive analyses of it. Wheatstone also was the first to draw the circuit in the distinctive diamond style used ever since.

The principle of the circuit is that if three resistances are known, and the current in the cross branch is zero, the fourth resistance can be calculated. The measurement can be made very accurately since zero current can be detected with extremely high accuracy using a sufficiently sensitive galvanometer.

This circuit is actually not used in bridge sensors, despite appearances; but it is a direct ancestor of the bridge sensor. Let's now look at the circuits used for measuring sensor resistance.

Bridge Measurement Systems

Resistance Measurement



$$R_g = R_1\left(\frac{V_E}{V_O} - 1\right)$$

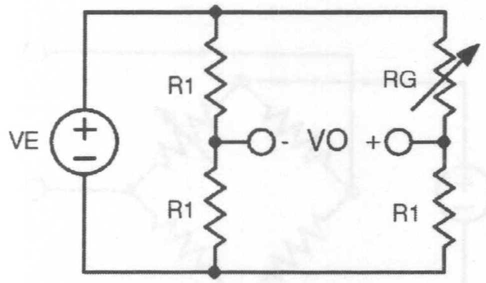
The circuit above shows how it is possible to measure resistance using a voltage divider. V_E is called the excitation voltage.

Note that the relationship between R_g and V_o is non-linear.

For a sensor like a strain gauge, this circuit will tend to produce very small changes in voltage, offset by a large amount. These are very difficult to measure if the offset is not known.

Bridge Measurement Systems

Removing the Offset



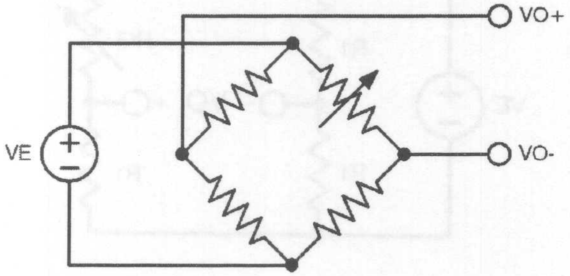
$$V_O = V_E\left[\frac{R_1}{R_1 + R_G} - \frac{1}{2}\right]$$

The large offset can be eliminated by adding a voltage divider and measuring the output differentially.

This assumes that R_g at rest is roughly equal to R_1, and that all the R_1s are very closely matched. Bridge sensors are nearly always built this way.

Bridge Measurement Systems

Single-Point Bridge



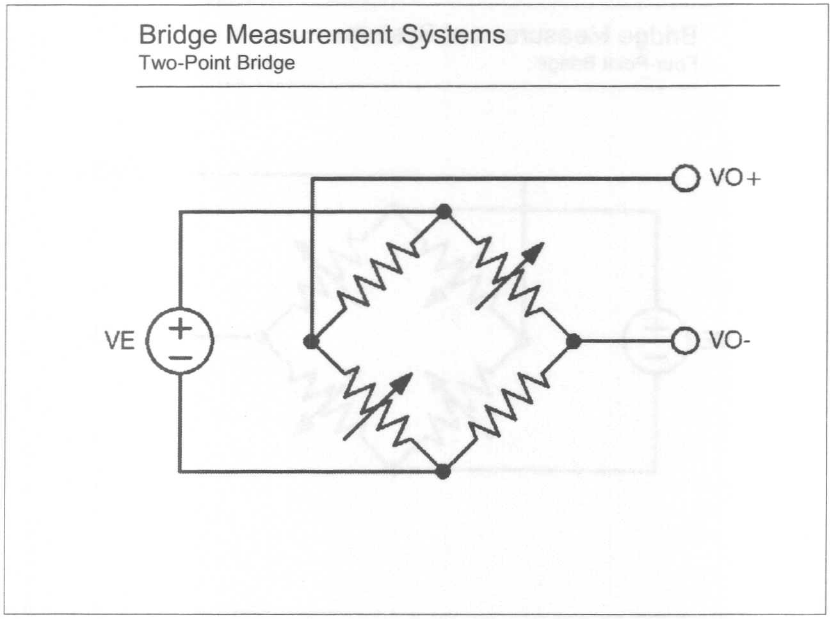
$$V_O = V_E\left[\frac{R_1}{R_1 + R_G} - \frac{1}{2}\right]$$

This circuit is electrically identical to the previous one. This is how bridge sensors are commonly drawn.

This configuration is called a single-point bridge sensor, since a single resistor changes, while the other three are fixed.

Again, the output voltage does not change linearly with the changing resistance.

Note that this circuit is not really a Wheatstone bridge, since the legs are not connected, and voltage is measured instead of current.



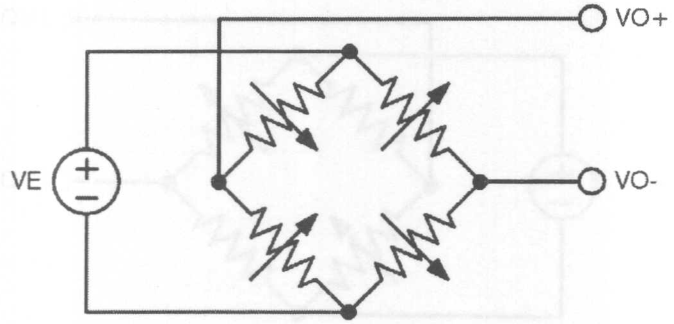
Bridge Measurement Systems

Two-Point Bridge

Bridge sensors can be made with two varying points. Higher-end load cells are commonly constructed this way, with two strain gauges connected oppositely. This makes the cell twice as sensitive.

Bridge Measurement Systems

Four-Point Bridge



Bridge sensors can also be made with all four elements varying. This is done in cases where a very linear and sensitive output is needed. Matching the elements can be difficult, so this is seldom done in inexpensive sensors.

Bridge Measurement Systems

Bridge sensor parameters

- **Sensitivity** – measured in mV/V; voltage output when the bridge is excited with 1 volt and the sensor is at full scale; 2mV/V is common
- **Accuracy** – measured in percent; load cells are typically 0.1% - 1% accurate, due to non-linear output curve
- **Bridge resistance** – resistance of the bridge measured unloaded across the signal terminals

Three parameters are usually supplied with bridge sensors.

Sensitivity is the most important. It is nearly always expressed in mV/V. A bridge with 2mV/V sensitivity will deliver 2mV if the measurement is at full-scale and the bridge is excited with 1V. This shows that the signals to be measured are very, very small; a 10V excitation can excite a 2mV/V sensor to at most 20mV!

Accuracy is generally rated at the factory. Although the 1% and 0.1% numbers look abysmal, in practice most sensors can perform better in controlled conditions, especially after calibration and linearization.

The bridge resistance indicates the size of the resistors in the bridge when the bridge is unloaded (idle). Bridge resistances are typically less than 1k. Most bridges measure the same resistance across the excitation terminals. Note that the bridge resistance determines power consumption, which can be rather high; a 10V excitation consumes 20mA for a 500-ohm bridge.

Bridge Measurement Systems

Bridge sensor error terms

- **Offset error** – voltage produced when the measurement parameter is zero
- **Full-scale error** – difference between the ideal voltage and the actual voltage when the measurement parameter is at maximum
- **Drift** – change of the above as temperature varies
- **Non-linearity** – deviation of the output curve from a straight line

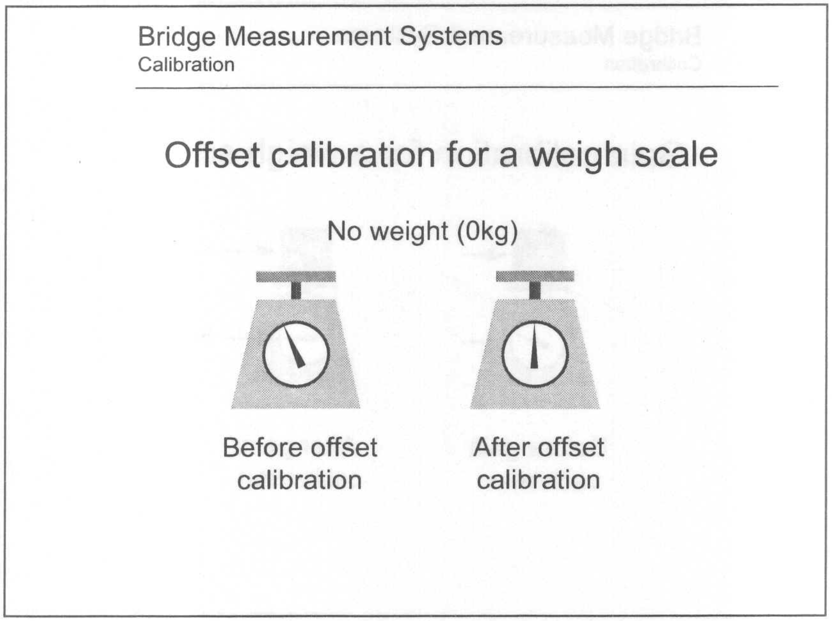
Bridge sensors have error terms similar to those found on other kinds of sensors.

Offset error is the voltage produced when the sensor's measurement parameter is zero. This can easily be calibrated out.

Full-scale error is the deviation in slope from the expected value. This can also be easily corrected.

Drift is very important: it is the change of the above values with temperature and time. It is much more difficult to compensate for drift – especially drift with time.

As we have seen, most bridge sensors exhibit an inherent non-linearity, in addition to the non-linearities arising from other sources. This can be corrected for to a degree, but commonly is not, as it is (with a good sensor) smaller than most other errors.



Bridge Measurement Systems
Calibration

Offset calibration for a weigh scale

No weight (0kg)

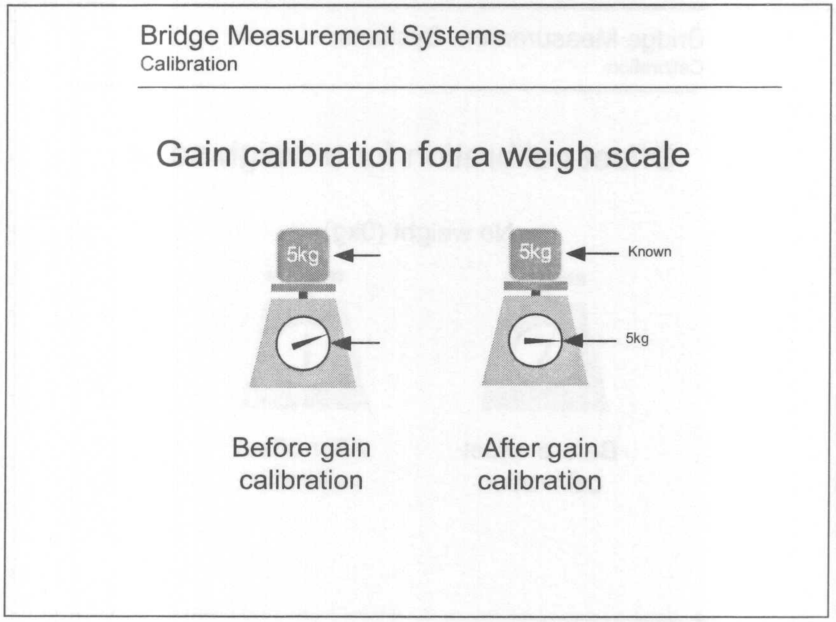
Before offset calibration

After offset calibration

This slide and the next illustrate calibration for a weigh scale.

In this slide, the scale is offset calibrated. Prior to calibration, the scale displays a negative offset. After calibration, with no weight applied, the scale reads zero, as it should.

(Note that in the case of an analog scale, calibration can be performed merely by rotating the movement of the gauge.)



Bridge Measurement Systems
Calibration

Gain calibration for a weigh scale

5kg
5kg
Known
5kg

Before gain calibration

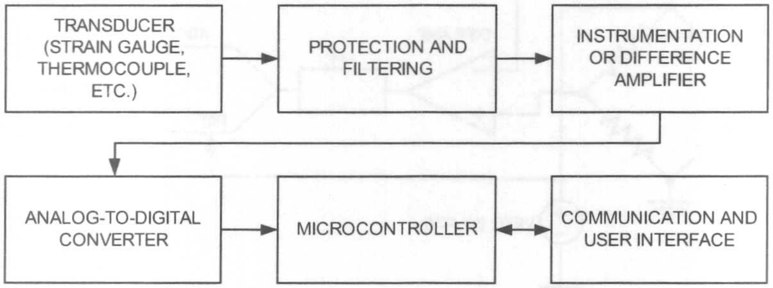
After gain calibration

In this slide, the scale is shown being calibrated for gain. A known weight is obtained and placed on the scale. The scale is then adjusted until its reading is equal to the mass of the known weight.

This scale reads too low when the 5kg known weight is placed on it. Following calibration, the scale reads 5kg with a 5kg known weight on it, as it should.

Bridge Measurement Systems

Typical bridge sensor system



TRANSDUCER (STRAIN GAUGE, THERMOCOUPLE, ETC.)

PROTECTION AND FILTERING

INSTRUMENTATION OR DIFFERENCE AMPLIFIER

ANALOG-TO-DIGITAL CONVERTER

MICROCONTROLLER

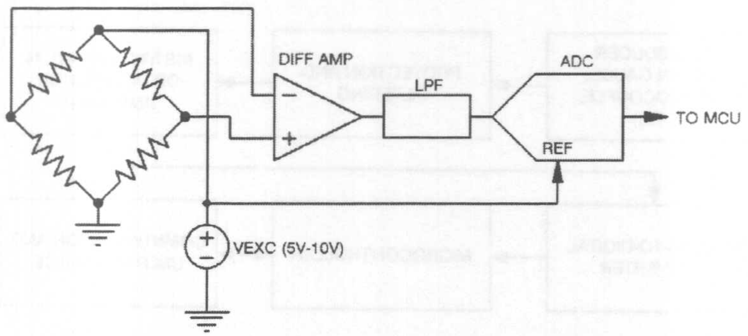
COMMUNICATION AND USER INTERFACE

A typical sensor measurement system has:

- An input stage, consisting of protection circuitry and amplification. In some circuits, the amplification is incorporated into the A/D converter.
- An A/D converter.
- A microcontroller, whose task is to collect data from the converter, possibly process the data collected, convert it to a useful form, and make it available to the user.

Bridge Measurement Systems

Basic connections: Amp + ADC

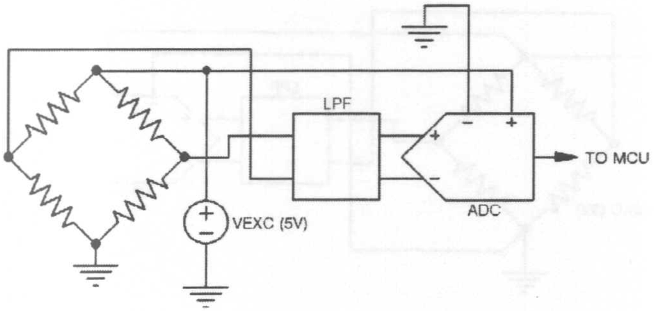


This is the most common circuit for bridge measurement using SAR converters. The signal is amplified using a differential amplifier of some kind, converting it from 0-20mV to the converter's input range, often 0-5V. This is then filtered.

This circuit shows a ratiometric connection. The excitation for the bridge is also used for the ADC's reference. In this connection, a four-wire load cell is used. We will see a six-wire load cell used shortly.

Bridge Measurement Systems

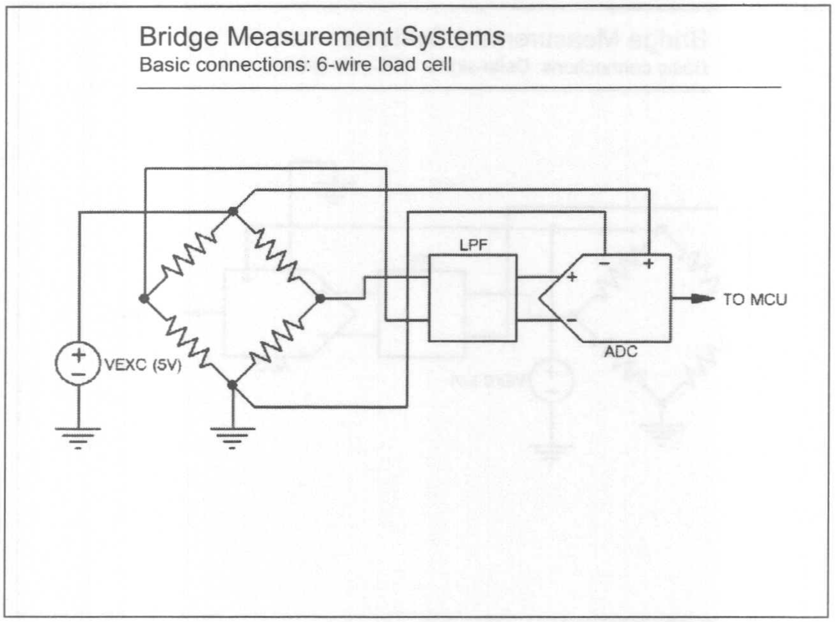
Basic connections: Delta-sigma ADC with gain



With a differential-input ADC having gain, as is the case with many industrial delta-sigma ADCs, the bridge sensor can often be connected directly. Here a differential low-pass filter is used for reduction of noise.

Again, this connection is ratiometric. Very few delta-sigma ADCs can accept larger than a 5V reference, so the excitation voltage is reduced to 5V. The reduction in gain is made up by the reduction in noise offered by a delta-sigma ADC.

Note that the gain in most delta-sigma converters is done by making adjustments to the modulator, and does not result from actual analog amplification. However, there are exceptions to this. TI's ADS1232 and ADS1234 are examples.



Bridge Measurement Systems

Basic connections: 6-wire load cell

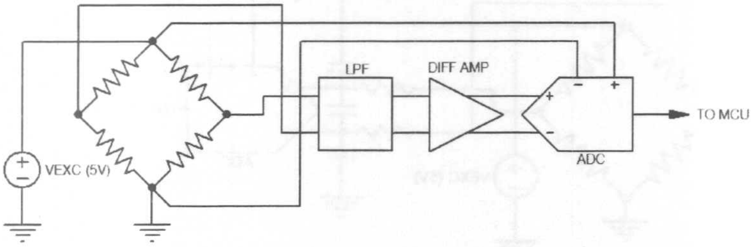
This schematic shows a 6-wire load cell connected in Kelvin fashion. For this to work, the ADC's reference input must have a very high impedance. If it does not, the reference must be buffered.

The Kelvin connection aims to reduce errors arising from voltage drops in the excitation line. The drops are caused by the relatively high current required to excite the load cell. Since the ADC's reference draws very little current, these drops do not occur in the return line, which is usually called the sense line.

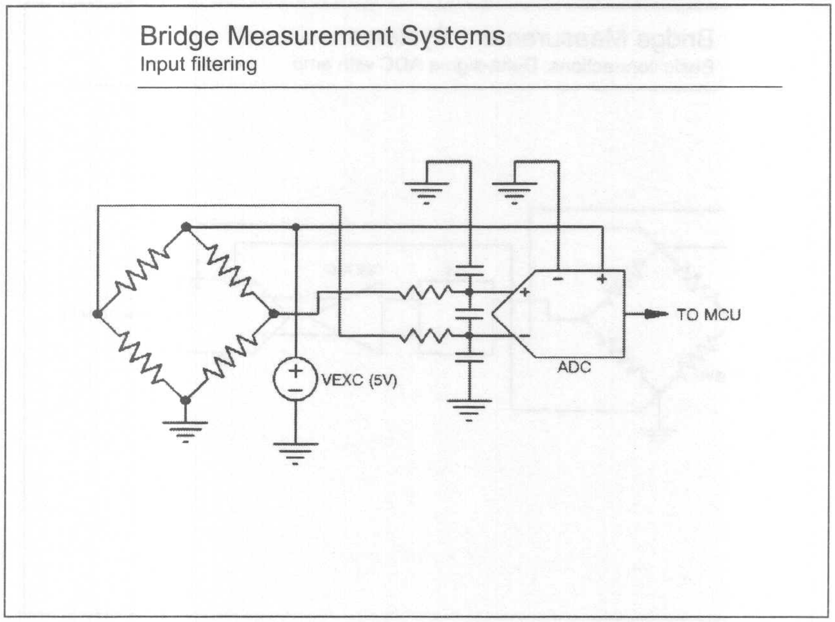
Kelvin connections are usually needed only for load cells connected through a long cable.

Bridge Measurement Systems

Basic connections: Delta-sigma ADC with amp



Although delta-sigma ADCs often have built-in amplification, sometimes greater performance can be obtained by adding an analog gain stage. This circuit shows this done differentially.



Bridge Measurement Systems

Input filtering

It's usually good to filter the output of a bridge sensor. This reduces wideband noise and can help reject interference.

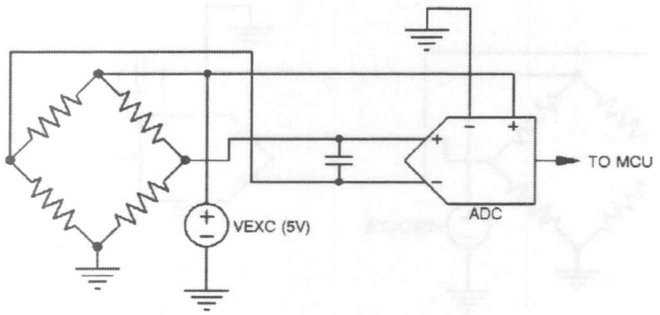
Shown is a single-pole differential low-pass filter configuration. The differential capacitor size depends on the desired data rate and settling time. It can be anywhere from 100pF to 1µF and above.

The common-mode capacitors should usually be very small, for example around 100pF. They are of limited benefit, but can reduce charge dumping from the input. They should be matched, or common-mode rejection will suffer.

The resistors should likewise be very small, no more than 100-200 ohms. If they are too large, increased noise and gain errors will result. They should also be matched for good common-mode rejection.

Bridge Measurement Systems

Better input filtering



Although most people think of the previously shown circuit as a necessary input filter, the simple one shown here can actually perform better.

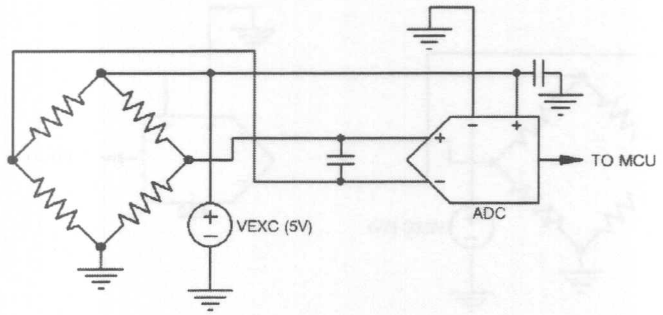
Since there are no components to be mismatched, common-mode rejection does not suffer. Eliminating the common-mode capacitors is of great benefit here.

Eliminating the resistors provides similar benefits. Note that there is some resistance already in the bridge itself. Also, many amplifiers and even ADCs have resistors built in to the front-end, so adding resistance usually only adds noise and gain error.

The differential capacitor is critical, and should be of very high quality. Ceramic X5Rs and worse should never be used. Even X7R-grade capacitors should be avoided. C0G or film capacitors should be used here.

Bridge Measurement Systems

Reference filtering



The lowly capacitor once again forms the most effective reference filter for delta-sigma converters.

Resistance can be of benefit, but more often is not, though it can be worth trying.

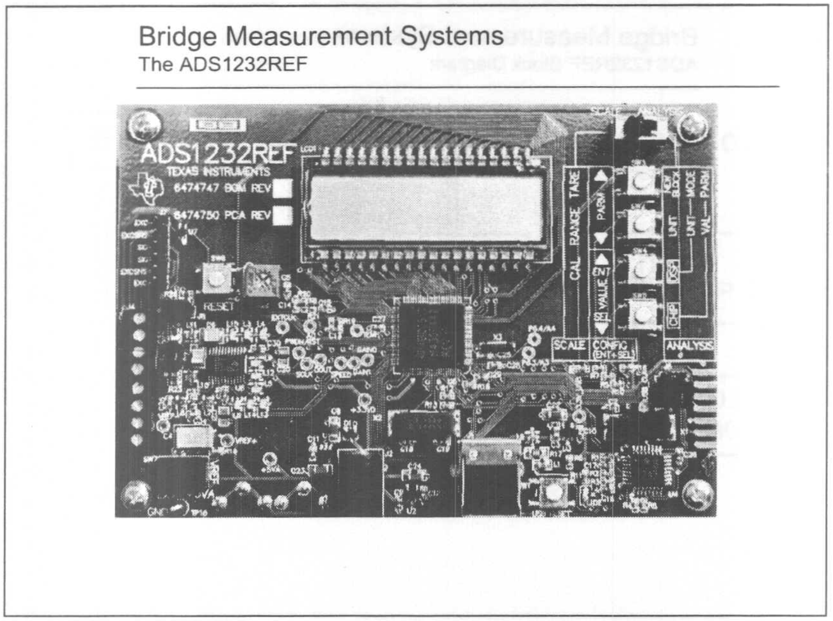
There is no one perfect capacitor size to use. Capacitor size should generally be as large as possible, though this depends on the reference source and system configuration. Different values should be tested, if possible, for effectiveness.

Note that if the reference is buffered by an op-amp, the op-amp's rated capacitive load typically does not apply, since the reference signal is essentially DC. It is possible to use 100µF and larger values without instability on nearly any op-amp.

Overly large capacitor values can cause differences between the voltage at the bridge excitation terminals and the reference voltage. It is wise to try different values.

A small capacitor should always be placed as close to the IC's reference pin as possible, for interference rejection.

The capacitor type is not as critical here. Tantalum and ceramic capacitors can be safely used. However, very low-quality capacitors should still be avoided, and if drift is a concern, higher-quality capacitors should be considered, as capacitors can cause reference drift.



Bridge Measurement Systems

The ADS1232REF

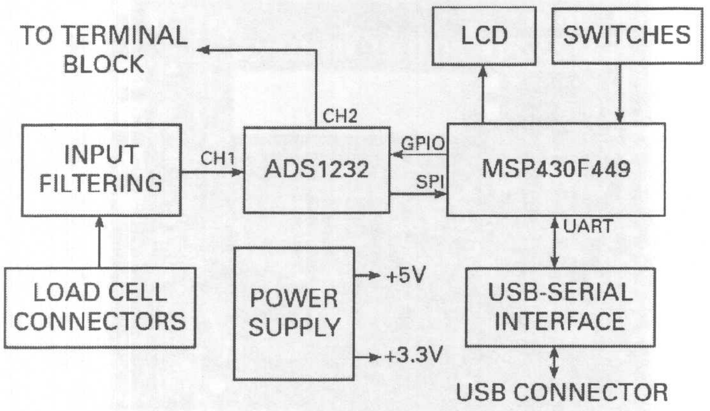
The ADS1232REF is a reference design for the ADS1232. It is designed to function as a weigh scale or load-cell meter, and has all of the components (except the case) which might be found in a real weigh-scale design.

It also serves as a general-purpose evaluation module for the ADS1232. It can be used either standalone or with a PC. The standalone modes display data from the board, and perform analysis on it. Software for the PC can read data from the board, perform analysis on it, and save it to disk.

The ADS1232REF can match and demonstrate all performance specifications shown in the datasheet for the ADS1232/34.

Bridge Measurement Systems

ADS1232REF Block Diagram

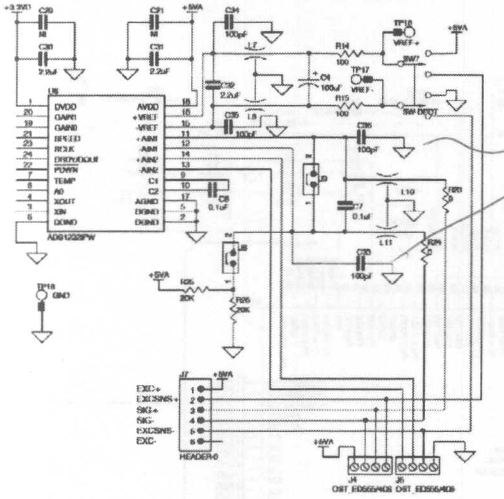


The ADS1232REF is based around an MSP430F449 microcontroller. This was chosen for its LCD driver, its multiplier, and its memory capacity.

The ADS1232 has three channels. Two are brought out to connectors, and the third reads an internal temperature diode. One of the channels is designed for connection to a load cell. One of the connectors has switchable excitation.

Bridge Measurement Systems

ADS1232REF ADC Section



FILTER
ISSUE

CAPS
ON
OSCILLATOR
~~[illegible]~~

The ADC section demonstrates several features of good design for the ADS1232, and a few extra features to aid evaluation.

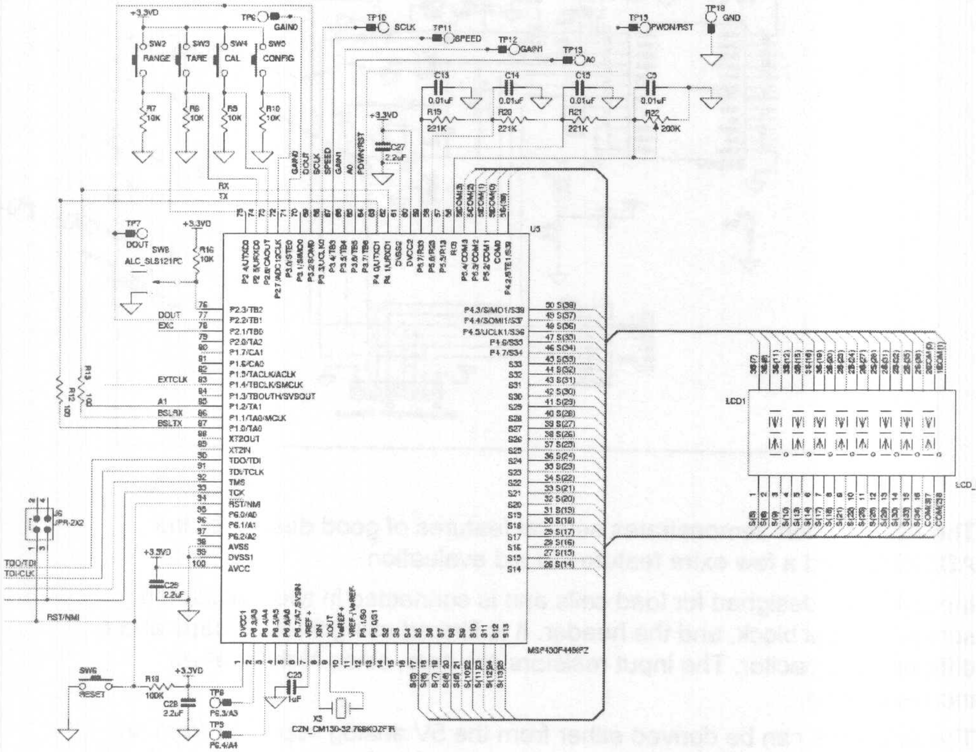
Input AIN1 is designed for load cells and is connected in two places; the screw-terminal block, and the header. It is filtered using through-caps and a differential capacitor. The input resistors are zero, since higher values increase noise.

The reference can be derived either from the 5V analog supply or from an external source – typically the excitation sense line from a six-wire load cell. The choice is made using switch SW7.

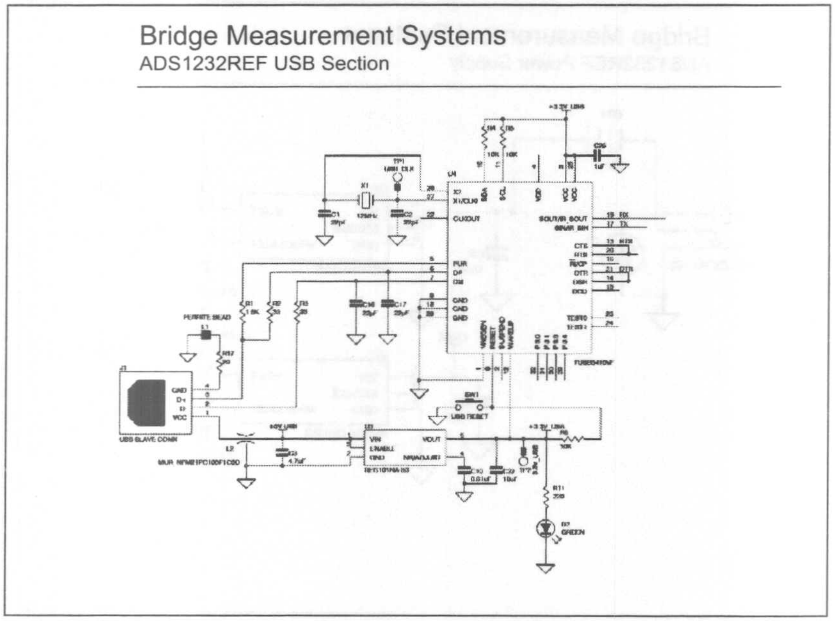
The reference is filtered by an elaborate network of capacitors and a pair of 100 ohm resistors. These help reject interference which may be picked up from a long cable.

Bridge Measurement Systems

ADS1232REF MCU Section



The MCU section is quite straightforward. An eight-digit starburst LCD occupies the LCD driver pins; five switches provide user interface. Programming comes from an (optional) JTAG connector or the serial bootstrap loader, which is connected to the USB to serial converter.



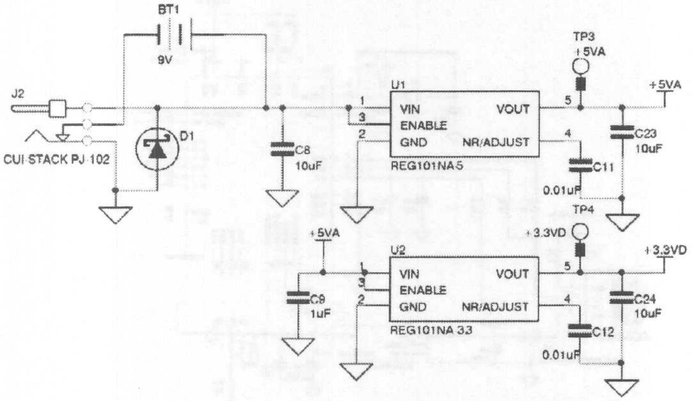
Bridge Measurement Systems

ADS1232REF USB Section

The USB-serial converter is based around the TI TUSB3410. TI provides "virtual COM port" drivers for several operating systems, including Windows. This section is powered from USB, and is not powered when the USB cable is removed.

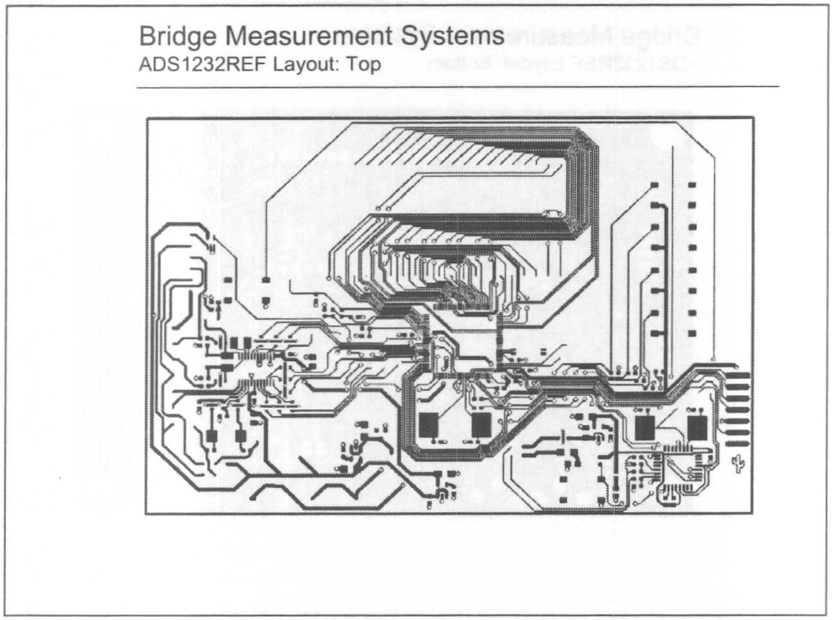
Bridge Measurement Systems

ADS1232REF Power Supply



Power is supplied either from a wall-mount AC adaptor or a 9V battery. A switched connector causes AC connection to override the battery. To save battery power, there are no power-on LEDs.

Low noise is important, since the power supply is (optionally) used as the ADC reference. The power supply is thoroughly filtered on both input and output.

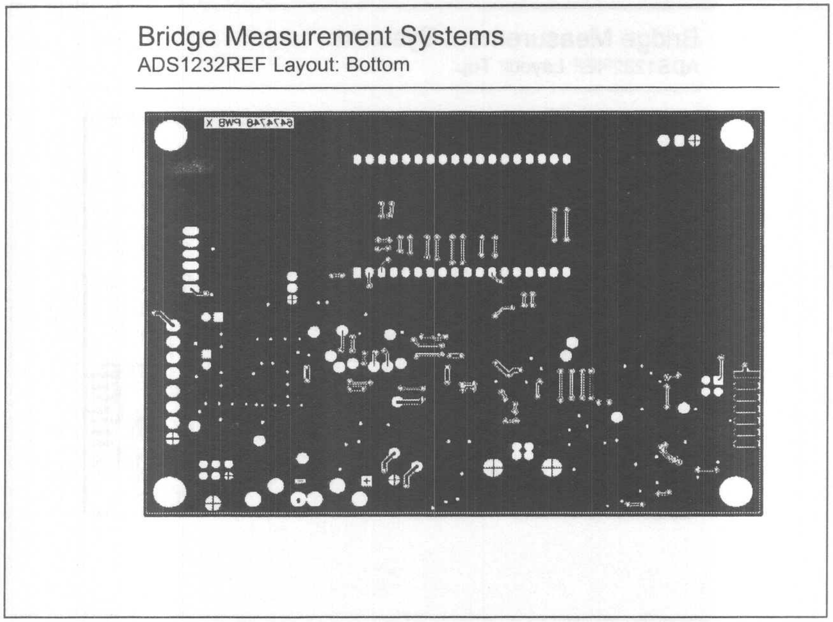


Bridge Measurement Systems

ADS1232REF Layout: Top

The ADS1232REF is laid out in two layers. This proves that good performance can be obtained with the ADS1232 even with a low-cost board process.

As many signals as possible are routed on the top layer, and the layout is arranged so that the microcontroller and fast digital components are kept away from the analog section.



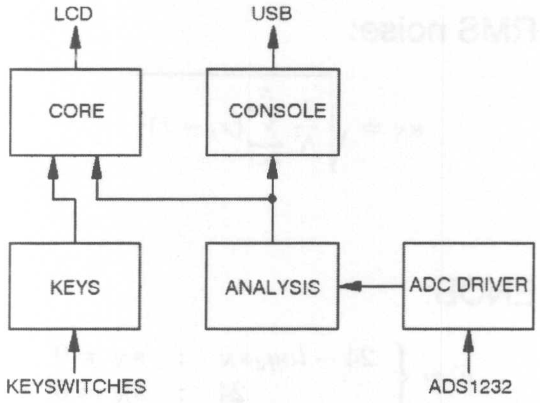
Bridge Measurement Systems

ADS1232REF Layout: Bottom

The bottom side serves as a ground plane. As few signal traces as possible are routed on the bottom side. Since the layout is not high-speed, wide, direct paths for return current are more important than a continuous ground plane.

Bridge Measurement Systems

ADS1232REF Firmware



The ADS1232REF firmware is written using a real-time operating system called FreeRTOS. FreeRTOS is designed for small microcontroller systems, and is very simple. It provides a kernel, process switching, a simple synchronization mechanism, inter-process communication using queues, timing functions, and rudimentary memory management.

The firmware uses four processes. The primary controlling process is CORE. This runs the display, and provides the board's standalone user interface. It obtains information on keyswitch events from KEYS, which scans and debounces the key switches, translating key presses into events. CORE obtains ADC data and block analysis results from ANALYSIS. ANALYSIS obtains ADC data from the ADC driver, which is not a process, but a collection of routines and an ISR.

The fourth process, CONSOLE, implements a serial command-line console with single-letter commands. The PC software uses the command line to communicate with the board. The CONSOLE interface can also be operated via a serial dumb terminal, through the virtual COM port driver for the TUSB3410. Like CORE, CONSOLE obtains data from ANALYSIS and can change certain ADC parameters.

Bridge Measurement Systems

Calculations

- RMS noise:

$$s_N = \sqrt{\frac{1}{N}\sum_{i=1}^{N}(x_i - \bar{x})^2}$$

- ENOB:

$$E = \begin{cases} 24 - log_2 s_N & : \quad s_N \neq 0 \\ 24 & : \quad s_N = 0 \end{cases}$$

The ADS1232REF firmware provides RMS noise and ENOB calculations.

To calculate RMS noise, the standard deviation formula is used. The result of this is codes, which is converted into volts if necessary.

For ENOB, the base 2 logarithm is taken on the result of the RMS noise calculation, and subtracted from 24. If the standard deviation is 0, 24 bits is used.

Bridge Measurement Systems

Summary

- Bridge sensors measure small changes in resistance using a pair of voltage dividers
- Delta-sigma ADCs are typically the best fit for bridge sensors
- Additional gain stages are optional with many delta-sigma ADCs
- Use ratiometric connections where possible
- Layout is important, but high-speed techniques are not usually needed

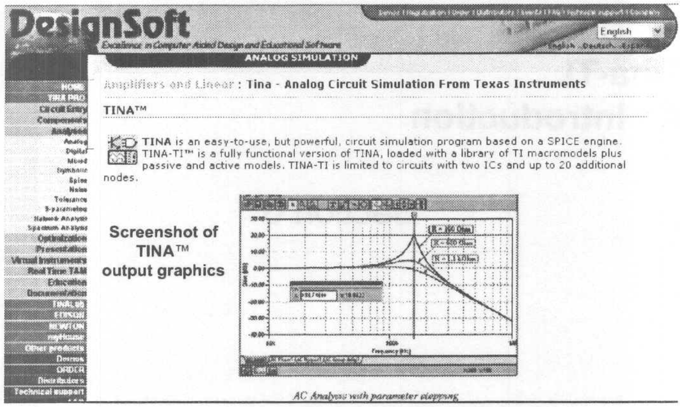
Tina-TI
An Introduction

Section 6

Tina-TI

An introduction

Tina-TI is a powerful analog circuit simulation tool



Amplifiers and Linear : **Tina - Analog Circuit Simulation From Texas Instruments**

TINA™

TINA is an easy-to-use, but powerful, circuit simulation program based on a SPICE engine. TINA-TI™ is a fully functional version of TINA, loaded with a library of TI macromodels plus passive and active models. TINA-TI is limited to circuits with two ICs and up to 20 additional nodes.

Screenshot of TINA™ output graphics

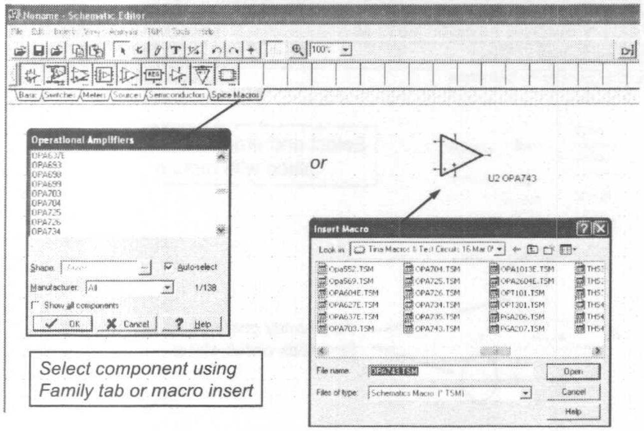
AC Analysis with parameter stepping

Tina-TI is a free, yet powerful, circuit simulator from DesignSoft and TI. This simulator accommodates up to 2 TI analog IC's and a wide pallet of passive and active devices. There is a maximum node count of 20. Note however, this count does not include the TI analog IC connections.

Tina-TI

An introduction

Tina-TI schematic editor



Select component using Family tab or macro insert

Select the circuit components including the TI analog IC's using the tabs located just below the component outlines. The tab names describe the family of components contained within the tab. Once a component has been selected it's just a mater of positioning it on the grid with the mouse.

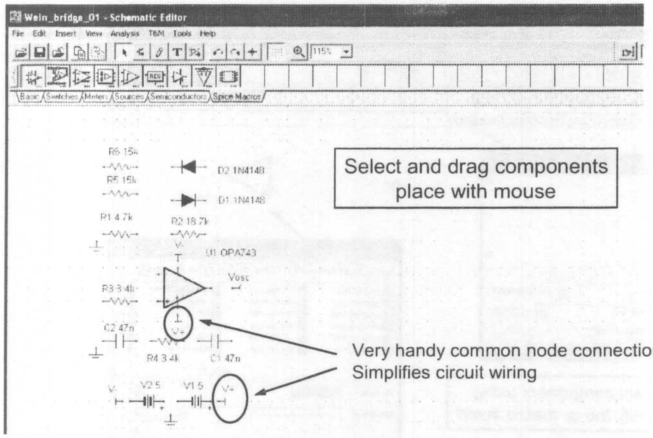
The "Spice Macro" tab accesses the library of TI analog devices: op-amps, difference amps, differential amplifiers, regulators, shunt current monitors, etc. Additional Pspice models can also be inserted in the circuit using the "macro insert" function found by clicking on the "Insert." In this example, the OPA743 op-amp was inserted using this macro function.

Double clicking on the other components opens a card listing the device characteristics. There are pull-downs on these cards that lets you select a specific device and alter certain characteristics.

Tina-TI

An introduction

Tina-TI schematic editor


Wein_bridge_01 - Schematic Editor
File Edit Insert View Analysis T&M Tools Help
Basic Switches Meters Sources Semiconductors Spice Macros
R6 15k
R5 15k
D2 1N4148
D1 1N4148
R1 4.7k
R2 18.7k
U1 OPA743
Vosc
R3 3.4k
C2 47n
R4 3.4k
C1 47n
V2 5
V1 5
V-
V+

Select and drag components place with mouse

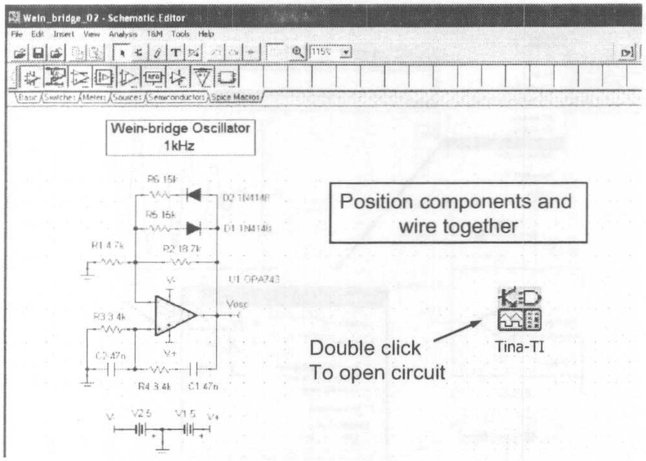
Very handy common node connection!
Simplifies circuit wiring

A wide variety of transistors, diodes, passive components, meters and sources can be selected from the various component tabs. Once selected they can be dragged and positioned as desired. Note the handy common node connection circuit element which can be used to simplify circuit wiring.

Tina-TI

An introduction

Tina-TI schematic editor


Wein_bridge_02 - Schematic Editor
File Edit Insert View Analysis T&M Tools Help
115%
Basic Switches Meters Sources Semiconductors Spice Macros
Wein-bridge Oscillator 1kHz
R6 15k
D2 1N4148
R5 15k
D1 1N4148
R1 4.7k
R2 18.7k
U1 OPA743
V-
Vosc
R3 3.4k
V+
C2 47n
R4 3.4k
C1 47n
V-
V2 5
V1 5
V+
Tina-TI

Position components and wire together

Double click
To open circuit

Once the components are placed they can be wired together. This is accomplished by positioning the hand symbol at one end of a component. Pressing the left mouse button starts the wire. Follow the grid with the wire to the next appropriate connection point. Once there, release the mouse button. The completed circuit is that of a 1kHz Wein-bridge oscillator.

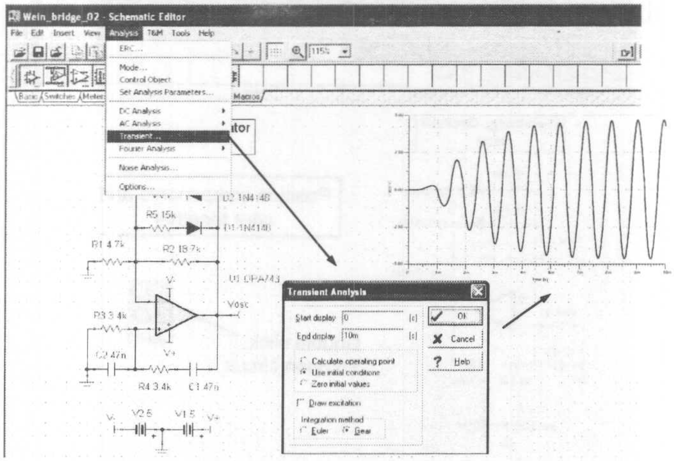
A connection check can be performed on the circuit using the ERC function. It is selected by clicking on "Analysis" pull-down. This may be skipped as an ERC will be first performed once any of the analysis options are selected.

Should an ERC error be detected, clicking on the listed error text will highlight the error on the schematic. This makes it much easier to locate and then correct the error.

Tina-TI

An introduction

Analysis Capabilities



Start display 0 [s]

End display 10m [s]

○ Calculate operating point
● Use initial conditions
○ Zero initial values

☐ Draw excitation

Integration method
○ Euler ● Gear

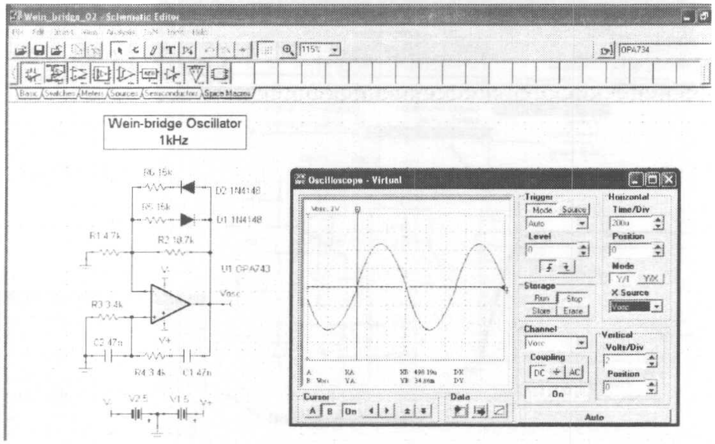
A number of analysis can be performed on the circuit; DC, AC, transient, Fourier and noise. There are more specific analysis available within these categories.

Shown here is a transient analysis made on the Wein-bridge oscillator circuit. The time scale has been selected to show the start-up and a few cycles of the oscillation.

Tina-TI

An introduction

Test and measurement capabilities



Wein_bridge_02 - Schematic Editor

Wein-bridge Oscillator
1kHz

Oscilloscope - Virtual

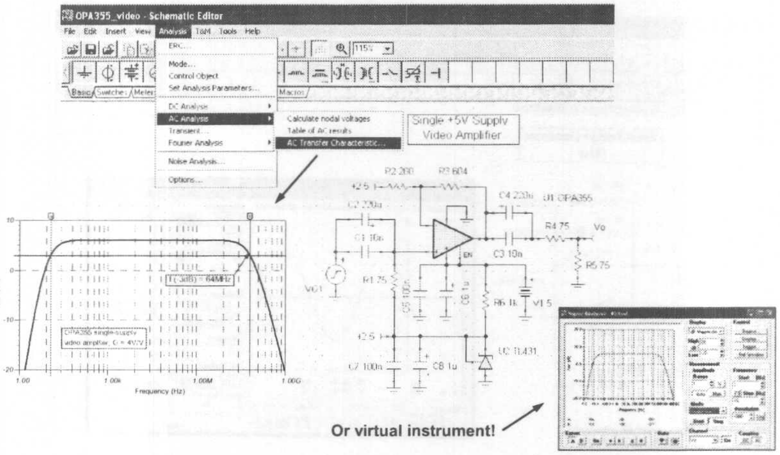
Trigger Mode: Auto
Level: 0
Time/Div: 200u
Position: 0
X Source: Vosc
Channel: Vosc
Volts/Div: 2
Position: 0
XB: 498.19u
YB: 34.89m

Tina-TI also provides virtual instruments. Here is an example of the virtual oscilloscope. Other instruments include; multimeters, signal generators, signal analyzers and x-y recorders.

Tina-TI

An introduction

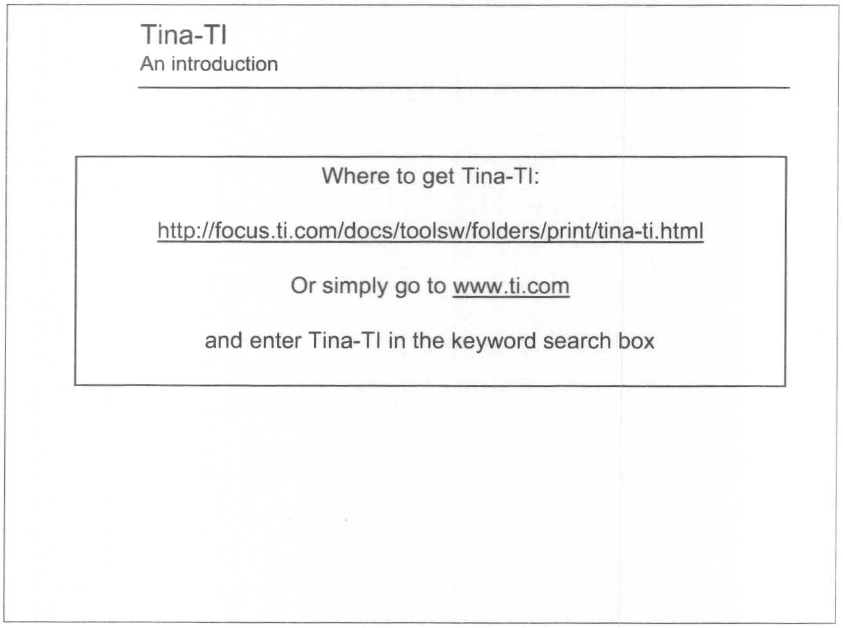
Test and measurement capabilities



Or virtual instrument!

Here's another example of Tina-TI's capability. A video amplifier application is shown using an OPA355 video op-amp and a TL431 shunt regulator IC. The TL431 provides a low impedance 2.5V voltage source. This is an example of a "2 TI device" circuit which is the limit for Tina-TI.

An amplitude vs. frequency plot can be acquired using the "Analysis" pull-down category "AC transfer characteristics." However, a virtual instrument can also be selected and displayed. Doing so allows the component values to be tweaked real time and the results immediately seen.



Tina-TI

An introduction

Where to get Tina-TI:

http://focus.ti.com/docs/toolsw/folders/print/tina-ti.html

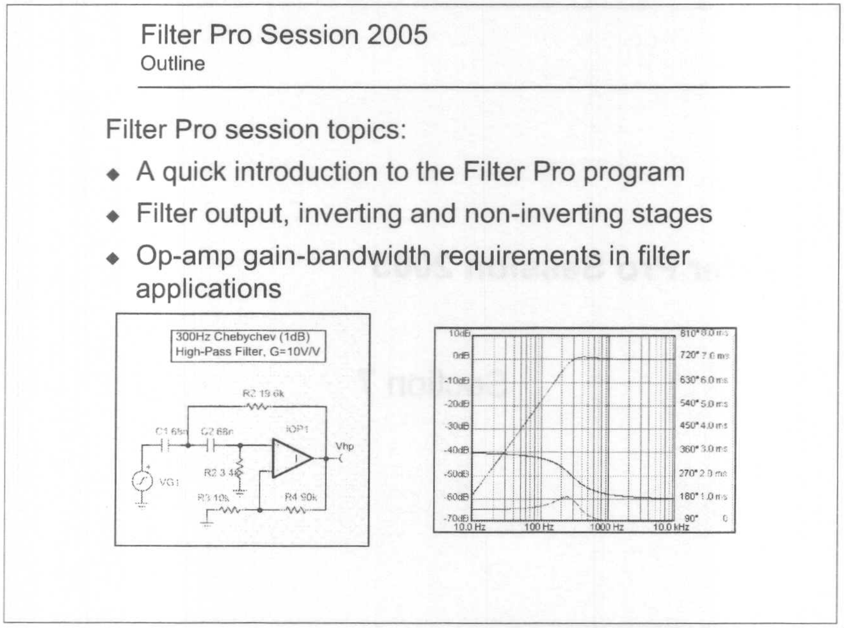
Or simply go to www.ti.com

and enter Tina-TI in the keyword search box

Tina-TI is available on the web and can be easily obtained via the TI website.

Filter Pro Session 2005

Section 7



Filter Pro Session 2005

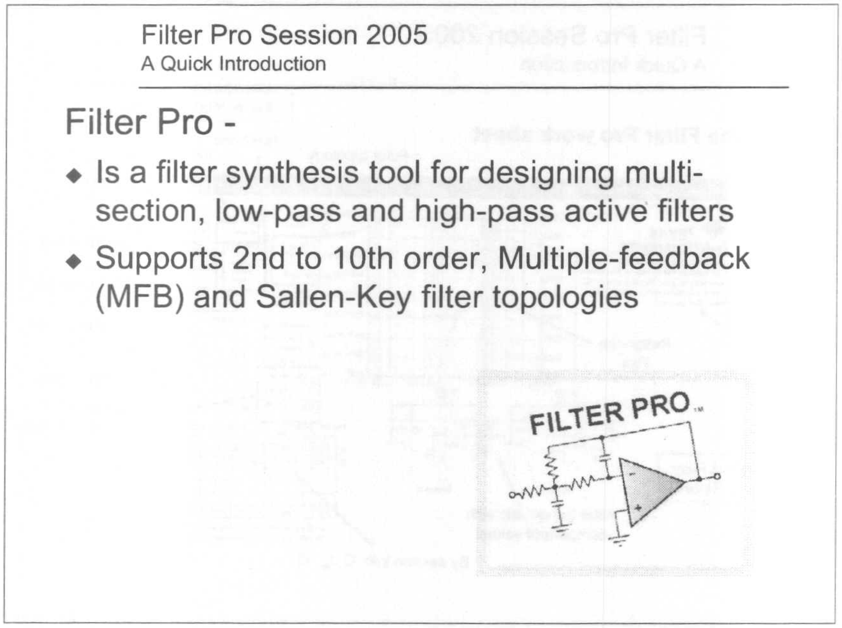
Outline

Filter Pro session topics:

- A quick introduction to the Filter Pro program
- Filter output, inverting and non-inverting stages
- Op-amp gain-bandwidth requirements in filter applications

300Hz Chebychev (1dB)
High-Pass Filter, G=10V/V
R2 19.6k
C1 68n
C2 68n
IOP1
Vhp
R2 3.4k
VG1
R3 10k
R4 90k
10dB
0dB
-10dB
-20dB
-30dB
-40dB
-50dB
-60dB
-70dB
810° 8.0 ms
720° 7.0 ms
630° 6.0 ms
540° 5.0 ms
450° 4.0 ms
360° 3.0 ms
270° 2.0 ms
180° 1.0 ms
90° 0
10.0 Hz
100 Hz
1000 Hz
10.0 kHz

This session provides a quick introduction to Filter Pro for those who are not familiar with the program, or a review for those who are. Then, two more advanced topics will be covered; filter output, inverting and non-inverting stages, and another on determining the gain-bandwidth requirements for op-amps used in filter applications.



Filter Pro Session 2005

A Quick Introduction

Filter Pro -

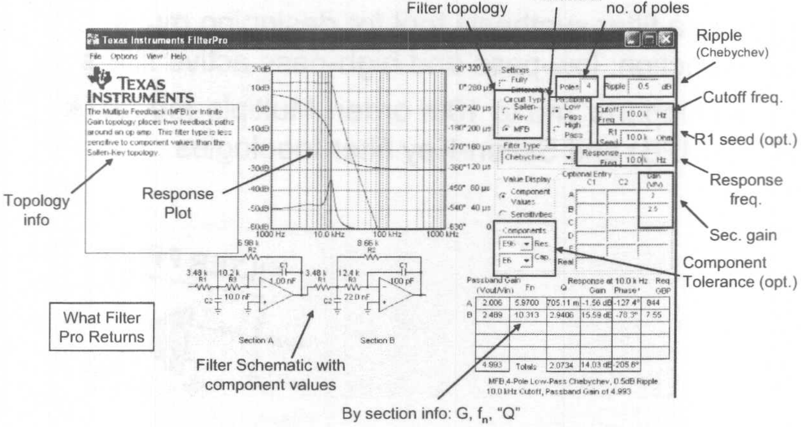
- Is a filter synthesis tool for designing multi-section, low-pass and high-pass active filters
- Supports 2nd to 10th order, Multiple-feedback (MFB) and Sallen-Key filter topologies

Simply stated, this is what Filter Pro provides the user.

Filter Pro Session 2005

A Quick Introduction

The Filter Pro work sheet



Filter entries you make

Filter topology

Passband

no. of poles

Ripple (Chebychev)

Cutoff freq.

R1 seed (opt.)

Response freq.

Sec. gain

Component Tolerance (opt.)

Topology info

Response Plot

What Filter Pro Returns

Filter Schematic with component values

By section info: G, f_n, "Q"

Texas Instruments FilterPro

The Multiple Feedback (MFB) or Infinite Gain topology places two feedback paths around an op amp. This filter type is less sensitive to component values than the Sallen-Key topology.

Circuit Type: ○ Sallen-Key ● MFB

Passband: ● Low Pass ○ High Pass

Poles 4

Ripple 0.5 dB

Cutoff Freq 10.0 k Hz

R1 Seed 10.0 k Ohm

Filter Type Chebychev

Response Freq 10.0 k Hz

Value Display: ● Component Values ○ Sensitivities

Components: E96 Res, E6 Cap

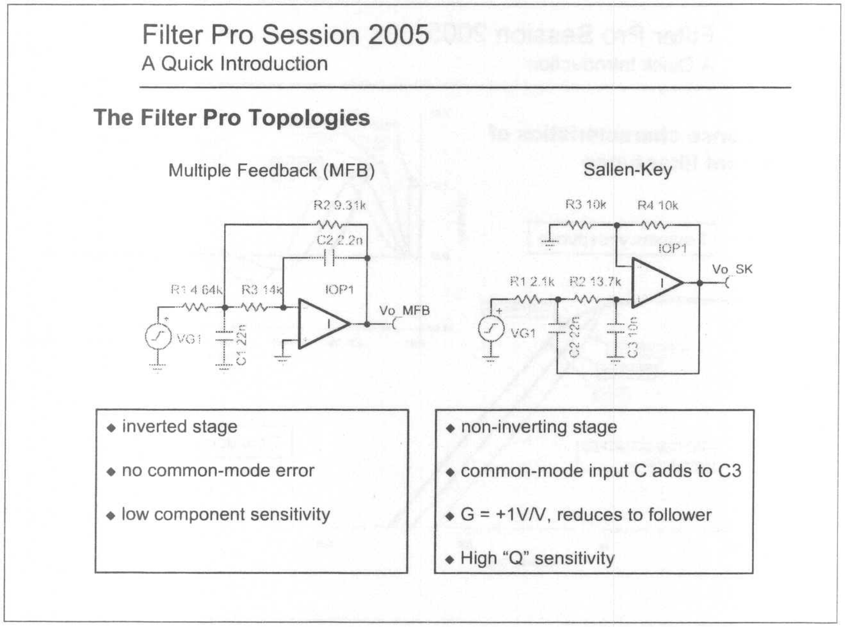
	Passband Gain (Vout/Vin)	Fn	Q	Response at 10.0 k Hz Gain	Phase°	Req GBP
A	2.006	5.9700	705.11 m	-1.56 dB	-127.4°	844
B	2.489	10.313	2.9406	15.59 dB	-78.3°	7.55
	4.993	Totals	2.0734	14.03 dB	205.8°	

MFB 4-Pole Low-Pass Chebychev, 0.5dB Ripple
10.0 kHz Cutoff, Passband Gain of 4.993

Once the general filter properties and requirements have been decided upon, they can be selected within, or entered in to the Filter Pro worksheet.

Filter Pro will provide a starter filter topology and component values. The response may be reviewed and the filter topology altered until the filter's electrical requirements are satisfied.

Upon the response requirements having been met, the topology and component values can be changed or iterated to best suit the user's needs.



Filter Pro Session 2005

A Quick Introduction

The Filter Pro Topologies

Multiple Feedback (MFB)

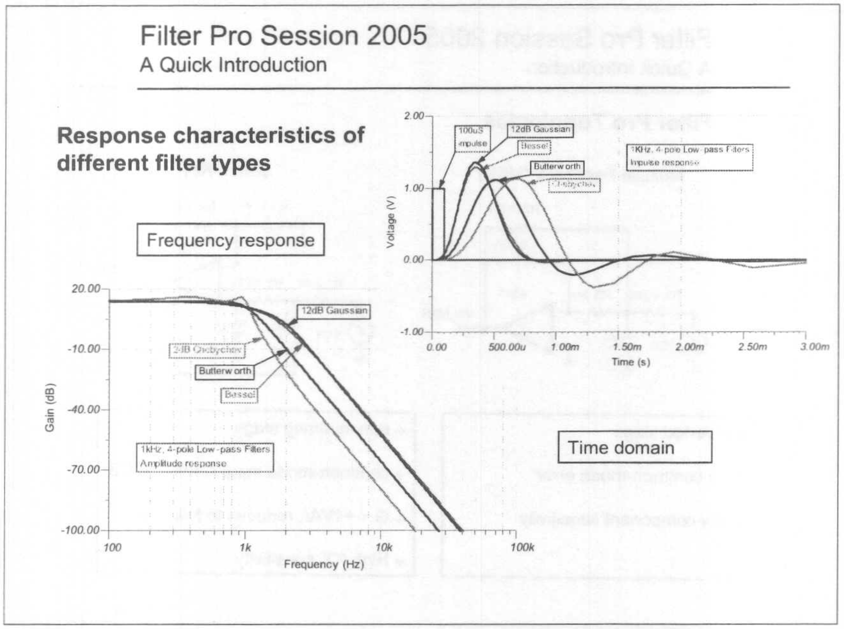
- inverted stage
- no common-mode error
- low component sensitivity

Sallen-Key

- non-inverting stage
- common-mode input C adds to C3
- G = +1V/V, reduces to follower
- High “Q” sensitivity

Filter Pro supports the Multiple Feedback (MFB) and Sallen-Key filter topologies:

- The MFB is a popular topology that features low sensitivity to component tolerances.
- Each 2nd-order MFB section produces an output phase inversion
- All Sallen-Key filters are non-inverting
- If the Sallen-Key gain is +1V/V the R3 is eliminated and R4 is 0Ω.
- The Sallen-Key does not show a high f_n sensitivity to component tolerance. However, the filter “Q” is very sensitive to component tolerance.



Filter Pro Session 2005

A Quick Introduction

Response characteristics of different filter types

Frequency response

Time domain

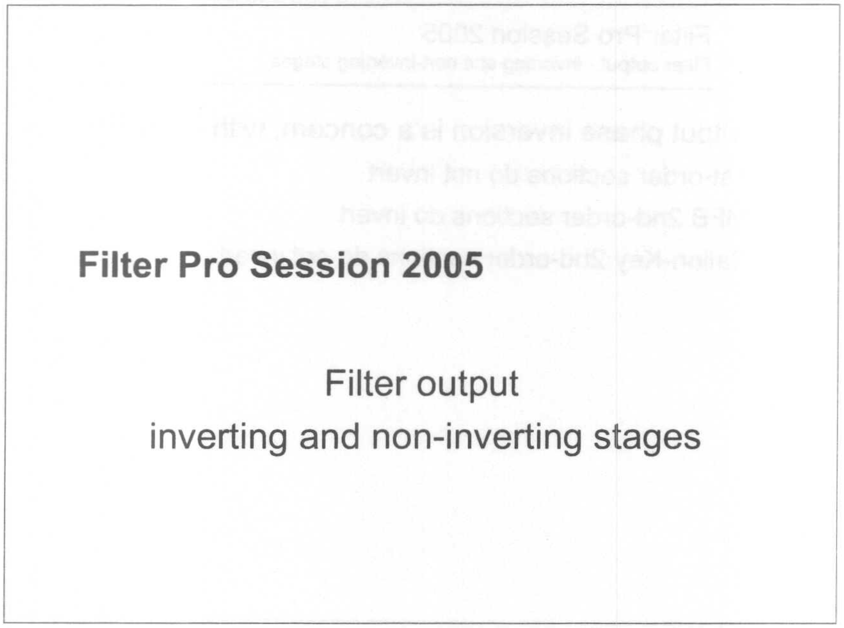
Presented here are frequency and time responses for 4 different filter types; Butterworth, Chebychev, Bessel and 12dB Gaussian low-pass filters.

◆ The Butterworth filter is noted for its maximally flat passband response and good pulse response.

◆ The Chebychev filter provides better attenuation beyond the passband, but the ringing in response to pulses may be objectionable.

◆ The Bessel filter also referred to as a linear phase filter has the best impulse response characteristics but the poorest rejection just beyond cutoff.

◆ The 12dB Gaussian filter results which follows a Gaussian distribution. This is most evident with bandpass filters. Its amplitude and pulse responses are similar to the Bessel filter.

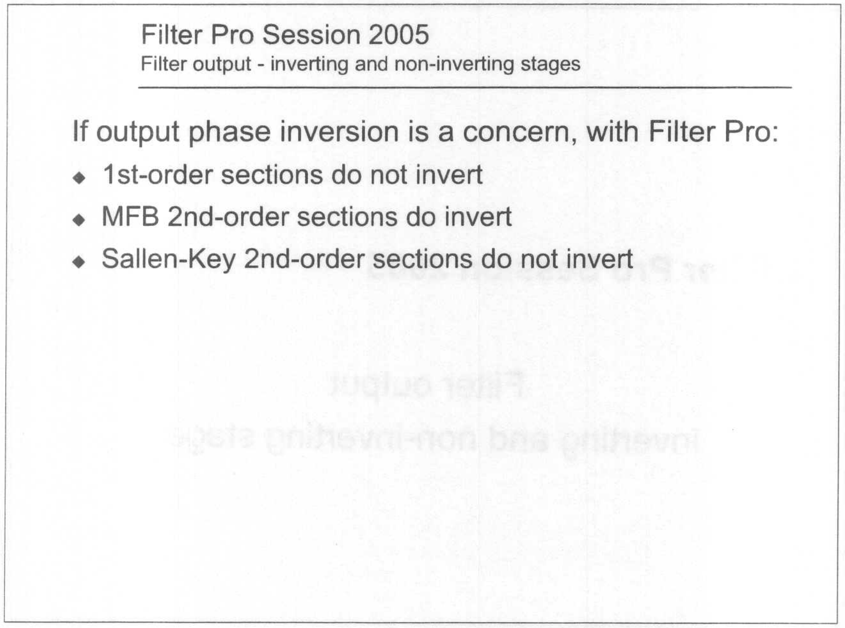


Filter Pro Session 2005

Filter output

inverting and non-inverting stages

Output phase inversion may be an issue to consider in some circuit applications. Filter Pro offers the choice of the Multiple Feedback (MFB) and Sallen-Key topologies. The filter topology that best suits the application, relative to inverted or non-inverted output issue, can be selected after considering their output characteristics.



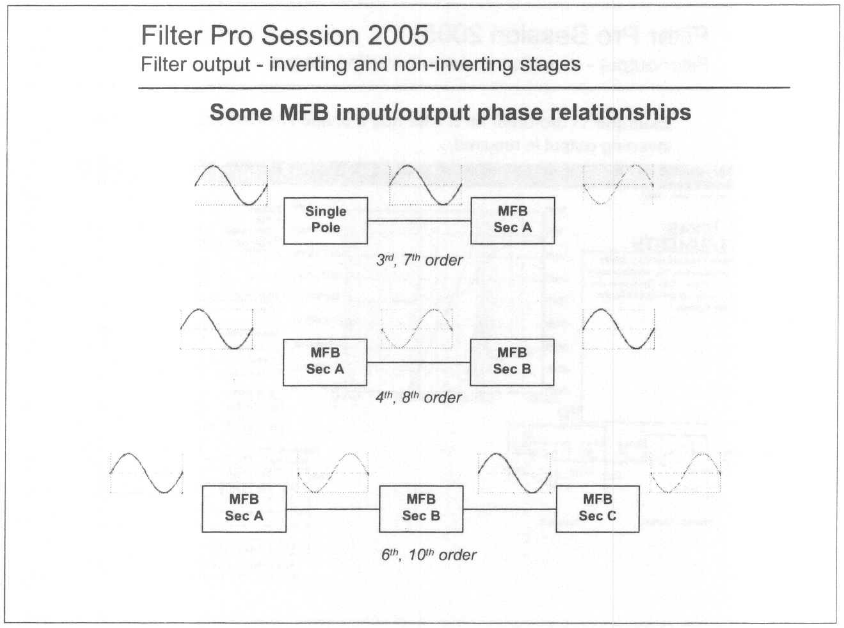
Filter Pro Session 2005

Filter output - inverting and non-inverting stages

If output phase inversion is a concern, with Filter Pro:

- 1st-order sections do not invert
- MFB 2nd-order sections do invert
- Sallen-Key 2nd-order sections do not invert

The phase inversion is the normal inversion associated with an inverting amplifier. Remember that filters introduce additional phase shift across frequency and that is in addition to the output inversion of the MFB section.



Filter Pro Session 2005

Filter output - inverting and non-inverting stages

Some MFB input/output phase relationships

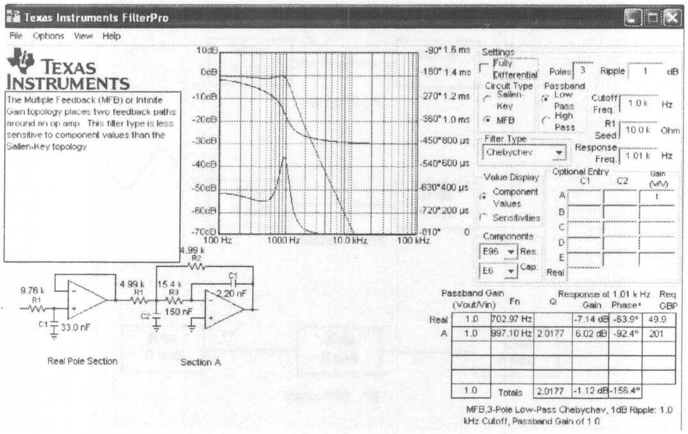
If using the MFB topology consider the following:

- Each 2nd-order MFB section introduces an output phase inversion
- A 3rd-order MFB filter inverts the output as well
- Adding 1 additional MFB section to a filter inverts the output
- Adding 2 sections returns the same output phase relationship as the preceding section

Filter Pro Session 2005

Filter output - inverting and non-inverting stages

Example: A 3rd-order MFB filter has a phase inverted output, but a non-inverting output is required…



The Multiple Feedback (MFB) or Infinite Gain topology places two feedback paths around an op amp. This filter type is less sensitive to component values than the Sallen-Key topology.

Settings
☐ Fully Differential
Poles: 3
Ripple: 1 dB
Circuit Type: ○ Sallen-Key ● MFB
Passband: ● Low Pass ○ High Pass
Cutoff Freq.: 1.0 k Hz
R1 Seed: 10.0 k Ohm
Filter Type: Chebychev
Response Freq.: 1.01 k Hz
Value Display: ● Component Values ○ Sensitivities
Components: E96 Res. E6 Cap.

Optional Entry

	C1	C2	Gain (V/V)
A			1
B			
C			
D			
E			
Real			

	Passband Gain (Vout/Vin)	Fn	Q	Response at 1.01 k Hz Gain	Phase°	Req GBP
Real	1.0	702.97 Hz		-7.14 dB	-63.9°	49.9
A	1.0	997.10 Hz	2.0177	6.02 dB	-92.4°	201
	1.0	Totals	2.0177	-1.12 dB	-156.4°	

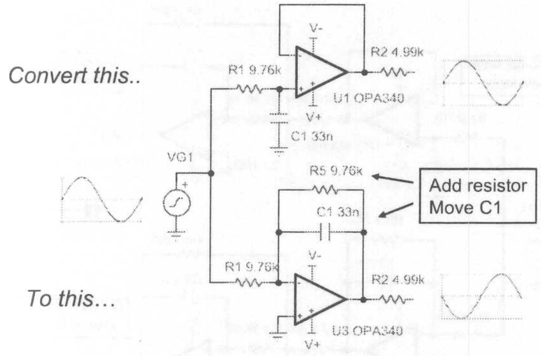
MFB,3-Pole Low-Pass Chebychev, 1dB Ripple: 1.0 kHz Cutoff, Passband Gain of 1.0

This application calls for a 3rd order Chebychev MFB low-pass filter, but with a non-inverting output. The non-inverting Sallen-Key would appear to be the logical choice but there could be "Q" sensitivity issues that would rule its use out.

Filter Pro Session 2005

Filter output - inverting and non-inverting stages

Real Pole Section



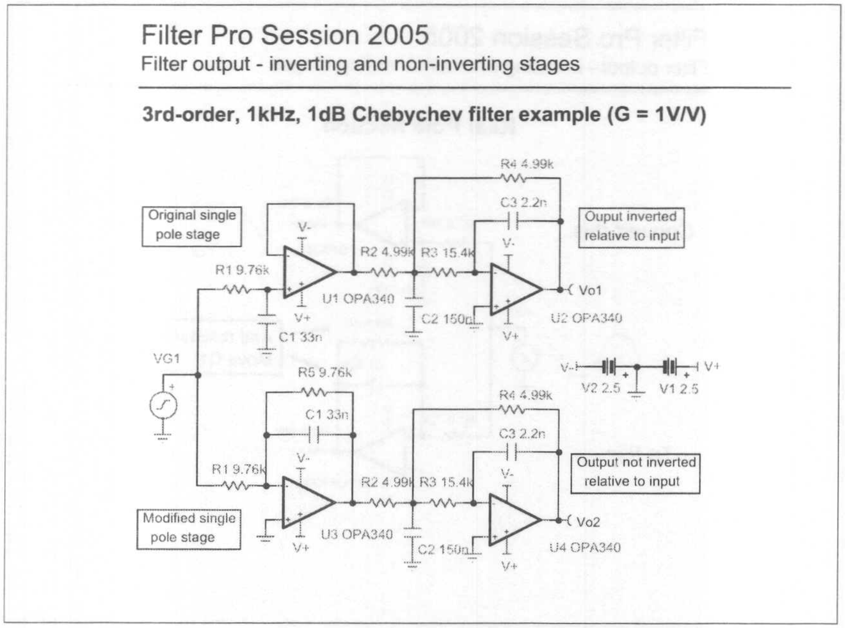
Convert this..

To this…

Add resistor
Move C1

The 1st -order, non-inverting section of the 3rd-order filter is shown in the schematic. It can be converted to an inverting stage by moving C1 as shown and adding a feedback resistor equal in value to the input resistor.

The signal inversion of this first stage, in conjunction with the 2nd-order stage that follows, results in the required input-output relationship.



Filter Pro Session 2005

Filter output - inverting and non-inverting stages

3rd-order, 1kHz, 1dB Chebychev filter example (G = 1V/V)

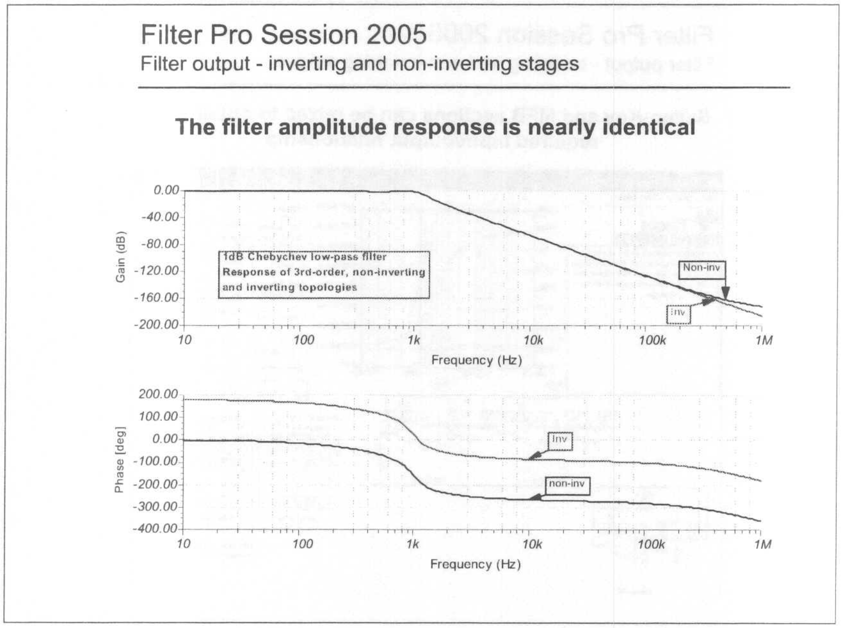
Original single pole stage

Ouput inverted relative to input

Modified single pole stage

Output not inverted relative to input

The upper half of the schematic shows the original 3rd order filter, which results in an inverted output. The lower half of the schematic shows the same filter with the first sectioned redesigned such that it inverts at its output. Since the 2nd-order stage that follows it also inverts, the two inversions result in a non-inverting output.



Filter Pro Session 2005

Filter output - inverting and non-inverting stages

The filter amplitude response is nearly identical

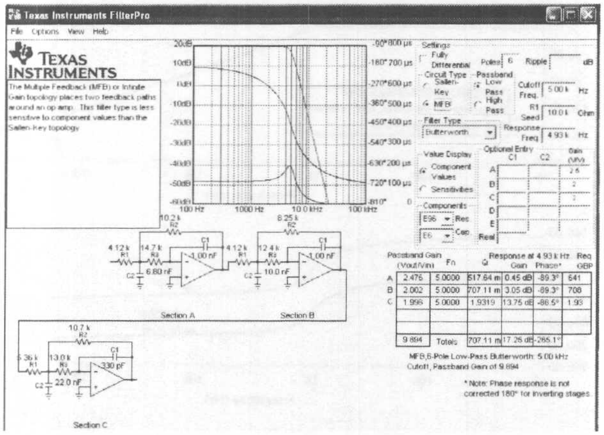
Here is a comparison of the 3rd-order filter's output before and after changing the 1st-order section from a non-inverting stage to an inverting stage. The overall output is now non-inverting. Note the 180 degree difference in the output phase responses.

A small amplitude deviation is noted starting around -160dB for the 3rd-order filter that has the non-inverting response. Likely, this would be of no consequence.

Filter Pro Session 2005

Filter output - inverting and non-inverting stages

Sallen-Key and MFB sections can be mixed to obtain the required input/output relationship



The Multiple Feedback (MFB) or Infinite Gain topology places two feedback paths around an op amp. This filter type is less sensitive to component values than the Sallen-Key topology.

Settings: ☐ Fully Differential; Poles 6; Ripple dB

Circuit Type: ○ Sallen-Key ● MFB

Passband: ● Low Pass ○ High Pass

Cutoff Freq. 5.00 k Hz

R1 Seed 10.0 k Ohm

Filter Type: Butterworth

Response Freq 4.93 k Hz

Value Display: ● Component Values ○ Sensitivities

Components: E96 Res; E6 Cap

Optional Entry	C1	C2	Gain (V/V)
A			2.6
B			2
C			2
D			
E			
Real			

	Passband Gain (Vout/Vin)	Fn	Q	Response at 4.93 k Hz Gain	Phase*	Req GBP
A	2.476	5.0000	517.64 m	0.45 dB	-89.3°	641
B	2.002	5.0000	707.11 m	3.05 dB	-89.3°	708
C	1.996	5.0000	1.9319	13.75 dB	-86.5°	1.93
	9.894	Totals	707.11 m	17.25 dB	-265.1°	

MFB,6-Pole Low-Pass Butterworth: 5.00 kHz Cutoff, Passband Gain of 9.894

* Note: Phase response is not corrected 180° for inverting stages.

This prototype, 3-section MFB, Butterworth low-pass filter has a gain of 10V/V and an f_c of 5kHz. One phase inversion can be eliminated by replacing a 2nd-order MFB stage with a Sallen-Key stage.

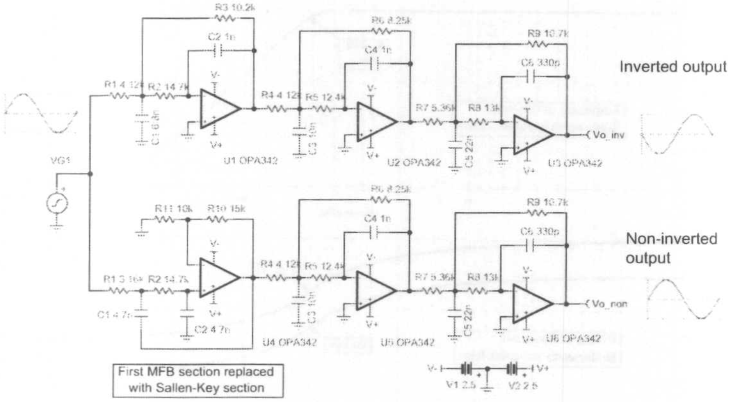
Since the gain, fn and Q are the same for each stage whether it is an MFB or Sallen-Key, Filter Pro can provide equivalent filter designs for both topologies. However, Filter Pro cannot simultaneously display the resulting filters for both topologies.

It is best to design the filter in one topology and then print the result. Then, select the other topology and let Filter Pro design it. Print that result. Since each stage in the two filters have identical responses, with the possible exception of inverting or non-inverting output, the corresponding stages can be plucked from one filter topology to the other.

Filter Pro Session 2005

Filter output - inverting and non-inverting stages

MFB 6th-order Butterworth low-pass filter

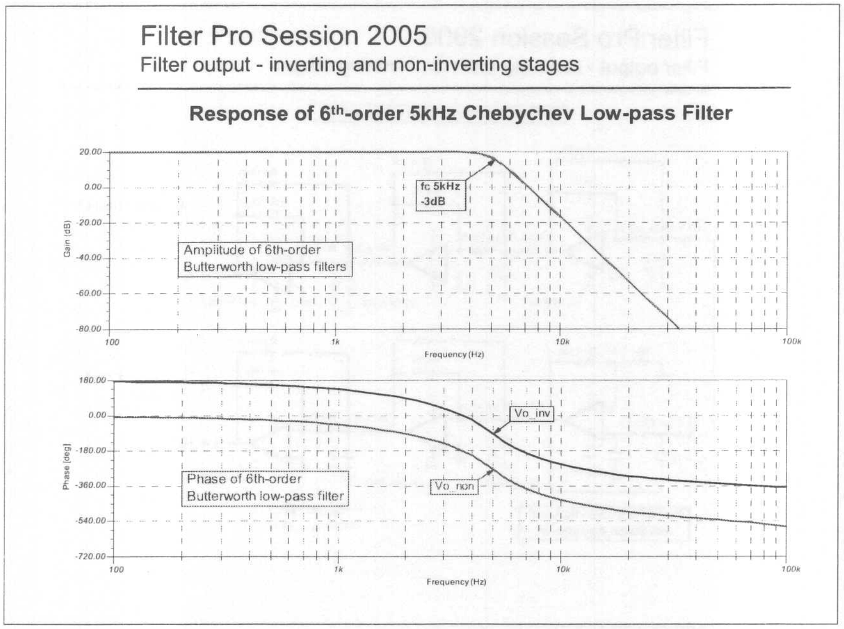


Inverted output

Non-inverted output

First MFB section replaced with Sallen-Key section

This is a TINA circuit representation for the 3-section, 6^{th}-order, low-pass filter in which the first MFB section has been replaced with a Sallen-Key section. The original MFB filter is shown in the top half, while the modified filter is shown in the lower half.



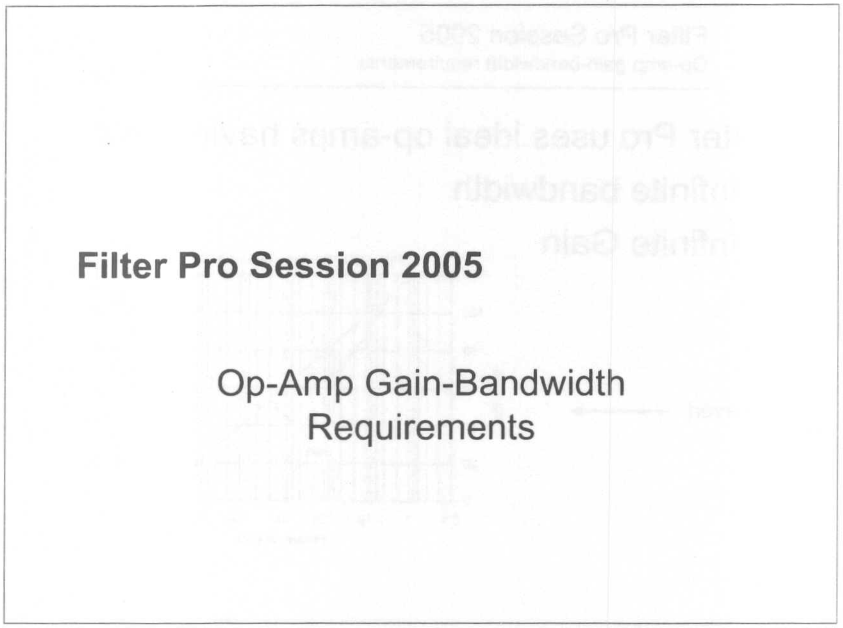
Filter Pro Session 2005

Filter output - inverting and non-inverting stages

Response of 6th-order 5kHz Chebychev Low-pass Filter

The TINA simulation provides the amplitude and phase response for the 3-section, 6th-order low-pass filter. Note the 180 degree phase difference for the 2 outputs.

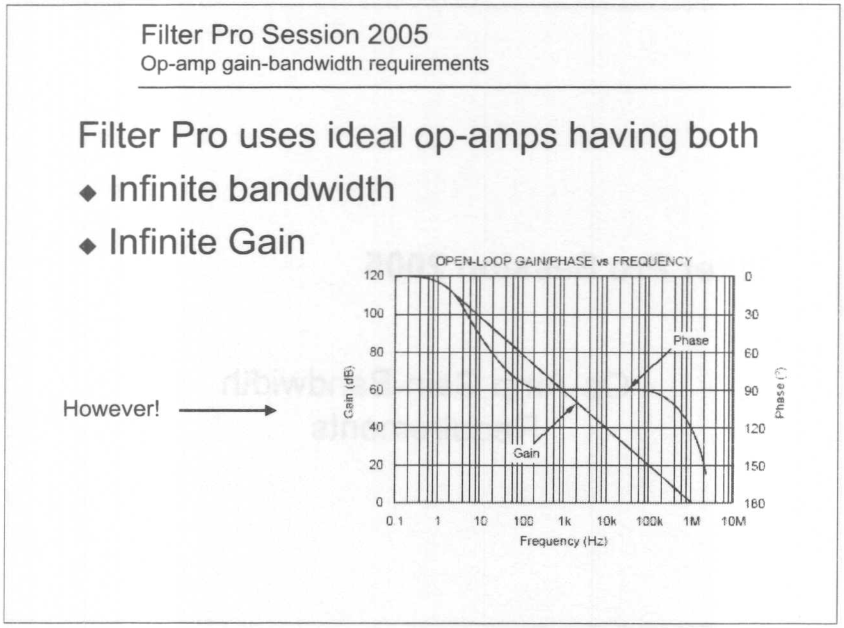
The amplitude response is identical for the standard MFB filter and the other MFB filter containing the Sallen-Key input section.



Filter Pro Session 2005

Op-Amp Gain-Bandwidth Requirements

It is not always intuitively obvious what the bandwidth requirements will be for the op-amps used in a particular filter. The requirements can be quite different from one filter response type to the next, even though they may have the same cut-off frequency.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

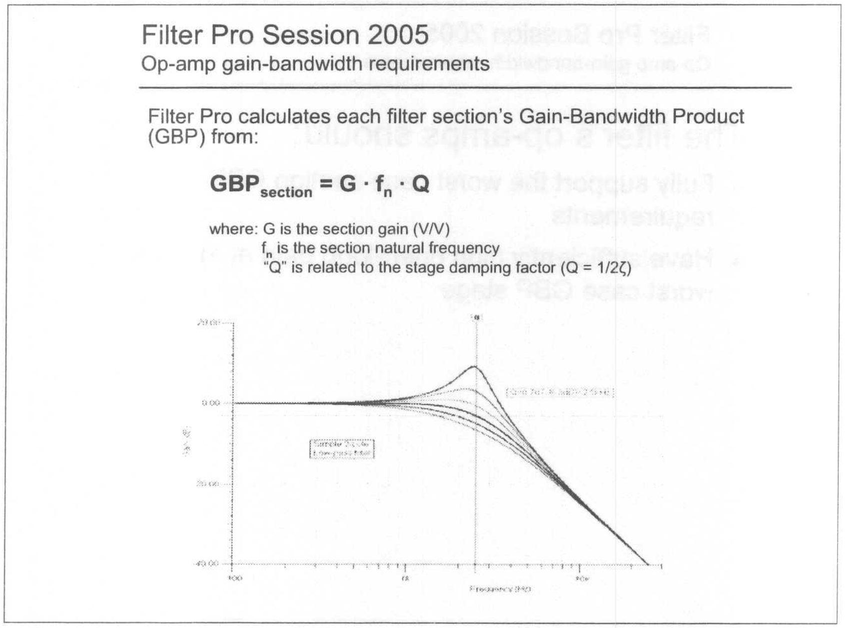
Filter Pro uses ideal op-amps having both

- ◆ Infinite bandwidth
- ◆ Infinite Gain

However! →

Filter Pro assumes the op-amps used in the filters it synthesizes are ideal. Bandwidth and gain are infinite resulting in mathematically perfect filter responses.

Real, non-ideal op-amps will produce far less than ideal response if they do not have sufficient gain and bandwidth to support the filter design criteria.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

Filter Pro calculates each filter section's Gain-Bandwidth Product (GBP) from:

$$\mathbf{GBP_{section} = G \cdot f_n \cdot Q}$$

where: G is the section gain (V/V)
f_n is the section natural frequency
"Q" is related to the stage damping factor ($Q = 1/2\zeta$)

Filter Pro provides a key indicator of the bandwidth required for each filter section. It is referred to as the Gain-Bandwidth Product (GBP). It consists of the product of the passband gain, the section's natural frequency and quality factor "Q." The GBP for each section is listed in the lower right table of the worksheet.

The "Q" is a damping constant determining the selectivity of the frequency response around f_n. Another way to define the "Q" is the ratio of the reactive power to the average power.

Filter Pro Session 2005

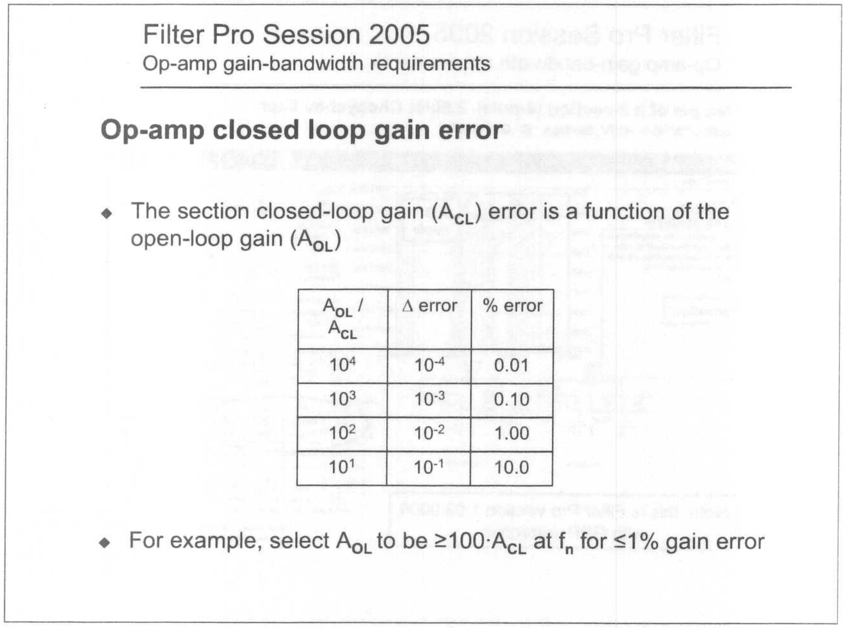
Op-amp gain-bandwidth requirements

The filter's op-amps should:

- Fully support the worst case section GBP requirements
- Have sufficiently high open-loop gain at fn of the worst case GBP stage

Real op-amps used to support the filter function are selected based on the worst-case section's GBP, such as, the section that has the highest GBP.

Although individual op-amps could be selected to meet the minimum requirements of each stage that isn't usually desired or even practical. Often, a dual or quad op-amp is used in the filter and one is selected that will meet the requirements of all the filter sections.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

Op-amp closed loop gain error

- The section closed-loop gain (A_{CL}) error is a function of the open-loop gain (A_{OL})

A_{OL} / A_{CL}	Δ error	% error
10^4	10^{-4}	0.01
10^3	10^{-3}	0.10
10^2	10^{-2}	1.00
10^1	10^{-1}	10.0

- For example, select A_{OL} to be $\geq 100 \cdot A_{CL}$ at f_n for $\leq 1\%$ gain error

The table shows the closed-loop gain error that results from limited open-loop gain. The particular application will dictate the acceptable maximum gain error.

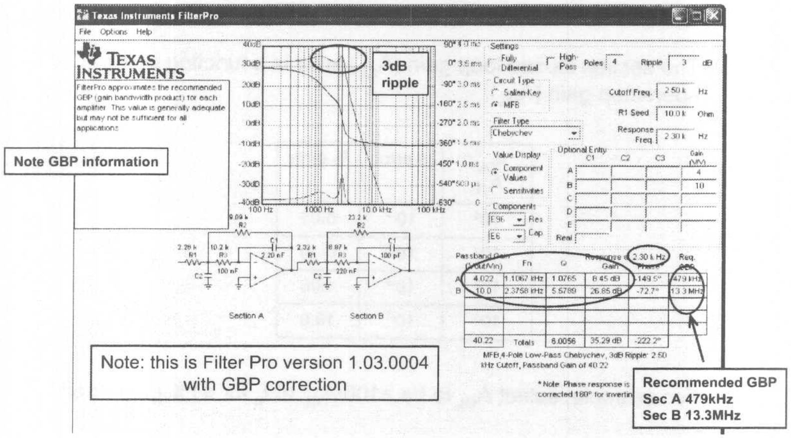
For example if a gain error of 1% is acceptable, then the ratio of open-loop gain (A_{ol}) to closed-loop gain (A_{cl}) should be no less than 100 at the worst-case f_n. Higher gain accuracy will require a higher A_{ol} to A_{cl} ratio.

Filter Pro Session 2005

Op-amp gain-bandwidth requirements

Example of a 2-section (4-pole), 2.5kHz Chebychev filter

(Section “A” G = -4V/V, Section “B” G = -10V/V)



3dB ripple

FilterPro approximates the recommended GBP (gain bandwidth product) for each amplifier. This value is generally adequate but may not be sufficient for all applications.

Note GBP information

Settings
☐ Fully Differential
☐ High Pass
Poles 4
Ripple 3 dB

Circuit Type
○ Sallen-Key
● MFB

Cutoff Freq. 2.50 k Hz
R1 Seed 10.0 k Ohm
Response Freq 2.30 k Hz

Filter Type
Chebychev

Value Display
● Component Values
○ Sensitivities

Components
E96 Res
E6 Cap

Optional Entry

	C1	C2	C3	Gain (V/V)
A				4
B				10
C				
D				
E				
Real				

	Passband Gain (Vout/Vin)	Fn	Q	Response at 2.30 k Hz Gain	Phase*	Req. GBP
A	4.022	1.1067 kHz	1.0765	8.45 dB	-149.5°	479 kHz
B	10.0	2.3758 kHz	5.5789	26.85 dB	-72.7°	13.3 MHz
	40.22	Totals	6.0056	35.29 dB	-222.2°	

MFB,4-Pole Low-Pass Chebychev, 3dB Ripple: 2.50 kHz Cutoff, Passband Gain of 40.22

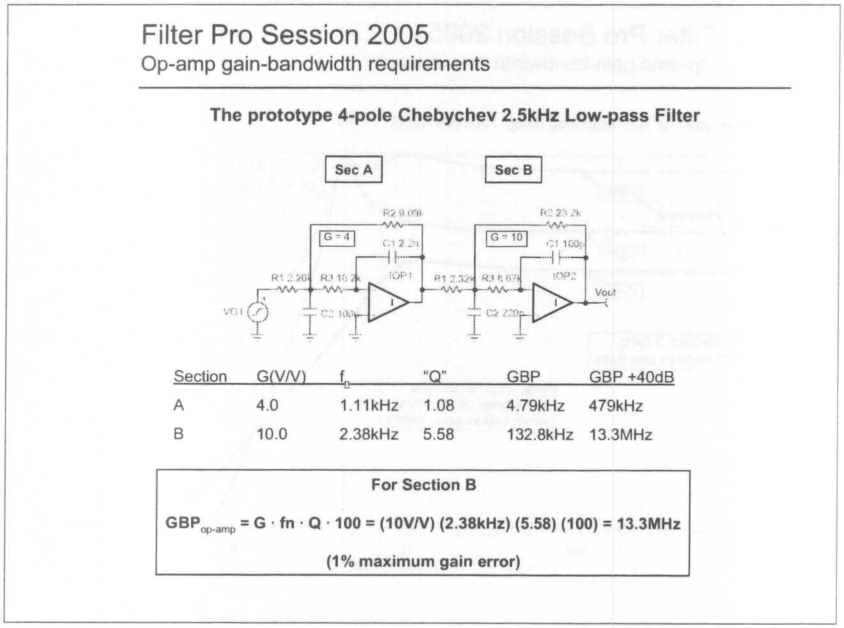
* Note: Phase response is corrected 180° for invertin

Note: this is Filter Pro version 1.03.0004 with GBP correction

Recommended GBP
Sec A 479kHz
Sec B 13.3MHz

Here is an example of a 2.5kHz Chebychev low-pass filter with a 3dB pass-band ripple and a gain of 40V/V. This filter places some pretty demanding requirements on the op-amps because of a high GBP section (section B). This example is used to illustrate how one goes about selecting an op-amp with sufficient bandwidth to support the requirements of this filter.

Filter sections that have the simultaneously requirements of high cutoff frequency, moderate to high voltage gain and a high “Q” will require a wide bandwidth amplifier.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

The prototype 4-pole Chebychev 2.5kHz Low-pass Filter

Section	G(V/V)	f_n	"Q"	GBP	GBP +40dB
A	4.0	1.11kHz	1.08	4.79kHz	479kHz
B	10.0	2.38kHz	5.58	132.8kHz	13.3MHz

For Section B

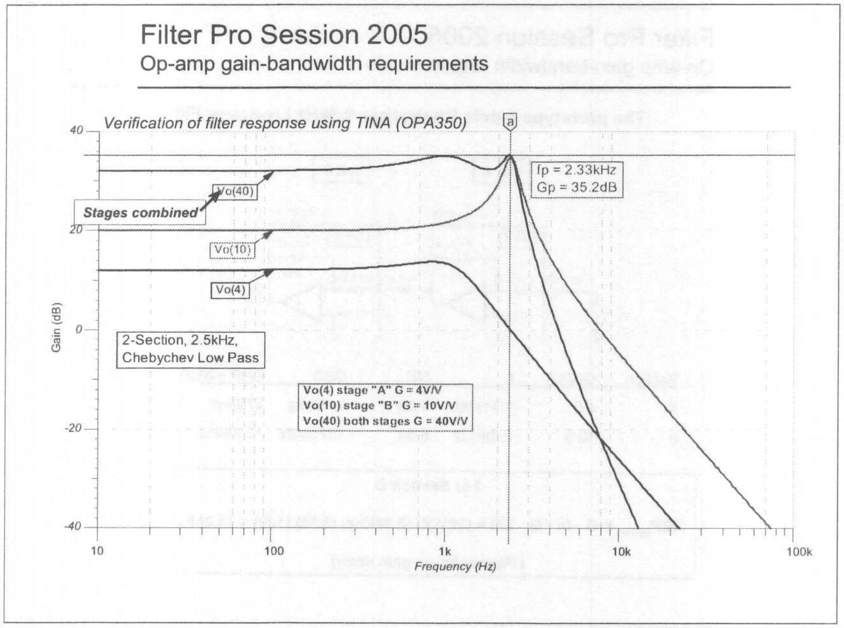
$GBP_{op\text{-}amp} = G \cdot fn \cdot Q \cdot 100 = (10V/V)\ (2.38kHz)\ (5.58)\ (100) = 13.3MHz$

(1% maximum gain error)

Filter Pro internally calculates a GBP of 4.8kHz for Section A, and 133kHz for Section B, and then multiplies each result by 100V/V (40dB). This gain is sufficient to assure a maximum gain error of 1%. A higher open-loop to closed-loop gain ratio would be required for a lower, maximum gain error.

Section B has the greater GBP requirement. Its required gain is the product of the nominal passband gain, 10V/V (20dB), the gain associated with the circuit "Q", 5.58V/V (14.9dB) at f_n and the aforementioned +40dB. Which results in a minimum section gain requirement of 75dB at f_n (Note that a 1% maximum gain error is again used).

$$GBP_{op\text{-}amp} = G \cdot f_n \cdot Q \cdot 100 = (10V/V)\ (2.38kHz)\ (5.58)\ (100) = 13.3MHz$$



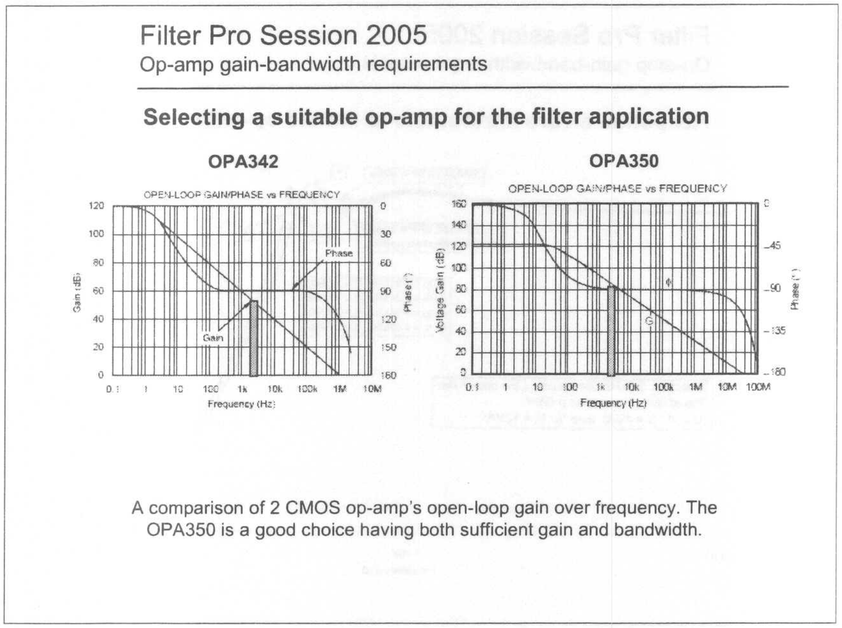
Filter Pro Session 2005

Op-amp gain-bandwidth requirements

This amplitude vs. frequency plot shows the simulated response of each section, A and B, and the overall filter output. The plot was obtained using the TINA simulation tool.

Filter Pro had calculated that the worst case GBP section was section B. It is labeled in the plot as Vo(10), indicating the stages G = 10V/V. Notice the response peaking at f_n due to the stage's higher "Q."

Combing section B's response with the response of section A results in the overall filter response.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

Selecting a suitable op-amp for the filter application

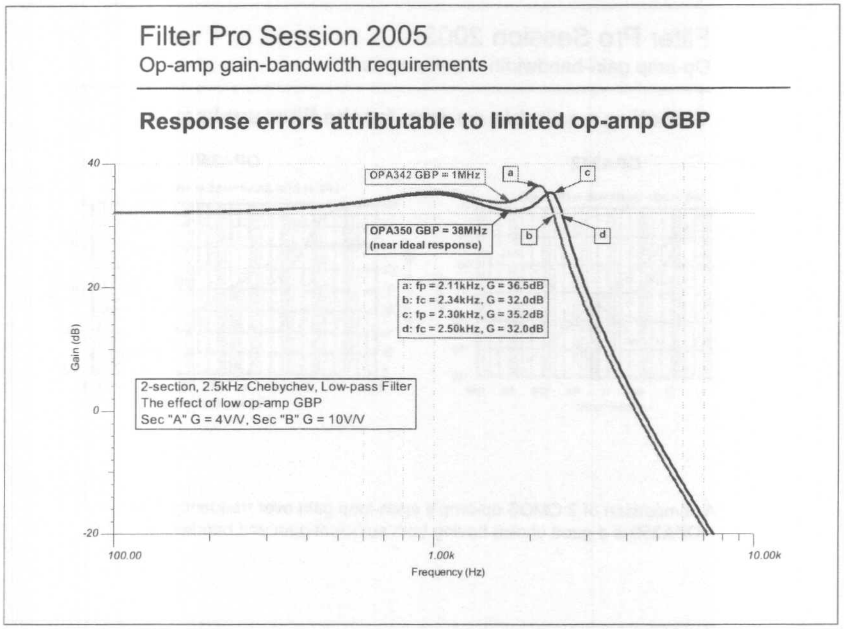
A comparison of 2 CMOS op-amp's open-loop gain over frequency. The OPA350 is a good choice having both sufficient gain and bandwidth.

Once the op-amp's GBP requirement is understood an appropriate op-amp can be selected. Here two different +5V CMOS op-amps are being considered for the 2.5kHz low-pass filter application; the OPA342 and OPA350.

A review of their open-loop gain (A_{ol}) vs. frequency plots shows the OPA342 exhibits approximately 50dB of A_{ol} at 2.5kHz, while the OPA350 has about 82dB A_{ol} at 2.5kHz.

Since section B requires about 75dB of A_{ol} at f_n, the OPA350 easily meets the minimum A_{ol} requirement; however, the OPA342 does not. The OPA350's unity gain crossover frequency of 30MHz, well exceeds the 13.3MHz unity gain bandwidth requirement.

Note that there may be more to selecting an op-amp than just its AC bandwidth performance. In comparison, the OPA350 has a quiescent operating current (IQ) of about 8mA, while the OPA340 IQ is only about 150µA. Then there are noise specifications which can be quite different for the two amplifiers. Broadband op-amps like the OPA350 use high-speed CMOS processing which is likely to exhibit much higher 1/f noise at low frequencies than an op-amp produced on a low speed process. It is wise to check all specifications in addition to the AC bandwidth.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

Response errors attributable to limited op-amp GBP

This is a gain vs. frequency plot for the filter with both the OPA342 and OPA350. Although the responses are similar in appearance for the two amplifier types, the OPA342 distorts the filter's cutoff frequency and gain.

The OPA350 produces a near ideal response. This was confirmed using ideal op-amps in the simulation.

Filter Pro Session 2005

Op-amp gain-bandwidth requirements

Section gain and GBP requirements

- The section gains can be iterated using Filter Pro to better match the GBP between sections

Chebychev, Fourth-order, 2.5kHz low-pass

Sec A V/V	Sec B V/V	GBP A* kHz	GBP B* kHz
40	1	47.8	13.3
20	2	23.9	26.6
10	4	11.9	53.3
8	5	9.61	66
5	8	5.92	107
4	10	4.79	133
2	20	2.39	266
1	40	1.19	533

* Doesn't include additional Aol/Acl gain

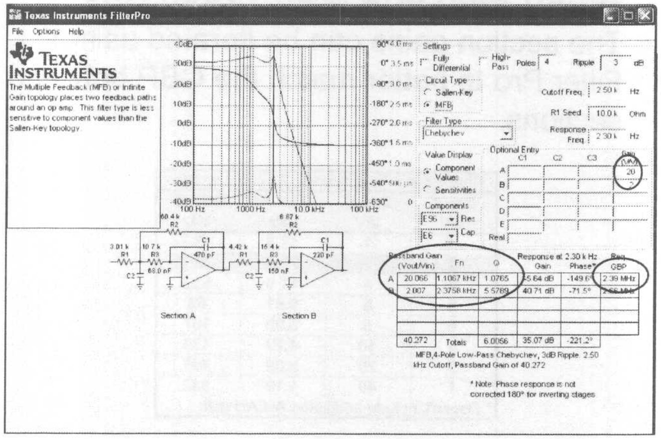
Setting the Section A gain to 20V/V and Section B to 2 V/V produces nearly identical GBP between sections. This gain set results in the lowest GBP of all combinations. Note that the GBP listed does not include the additional gain needed to establish a maximum gain error, such as, x100, x1000, etc.

Admittedly, iterating successfully becomes more difficult as the number of filter sections increases; however, an acceptable compromise can usually be reached.

Filter Pro Session 2005

Op-amp gain-bandwidth requirements

The gain has been redistributed to better match the GBP between sections



Texas Instruments FilterPro

File Options Help

The Multiple Feedback (MFB) or Infinite Gain topology places two feedback paths around an op amp. This filter type is less sensitive to component values than the Sallen-Key topology.

Settings: ☐ Fully Differential ☐ High-Pass Poles 4 Ripple 3 dB

Circuit Type: ○ Sallen-Key ● MFB

Filter Type: Chebychev

Cutoff Freq. 2.50 k Hz

R1 Seed 10.0 k Ohm

Response Freq. 2.30 k Hz

Value Display: ● Component Values ○ Sensitivities

Components: E96 Res, E6 Cap

Optional Entry

	C1	C2	C3	Gain (V/V)
A				20
B				2
C				
D				
E				
Real				

	Passband Gain (Vout/Vin)	Fn	Q	Response at 2.30 k Hz Gain	Phase*	Req. GBP
A	20.066	1.1067 kHz	1.0765	-5.64 dB	-149.6°	2.39 MHz
B	2.007	2.3758 kHz	5.5789	40.71 dB	-71.5°	2.66 MHz
	40.272	Totals	6.0056	35.07 dB	-221.2°	

MFB,4-Pole Low-Pass Chebychev, 3dB Ripple. 2.50 kHz Cutoff, Passband Gain of 40.272

* Note: Phase response is not corrected 180° for inverting stages

Filter Pro confirms that setting section A gain to 20V/V and section B to 2V/V results in nearly identical GBP for both sections.

Filter Pro Session 2005

Op-amp gain-bandwidth requirements

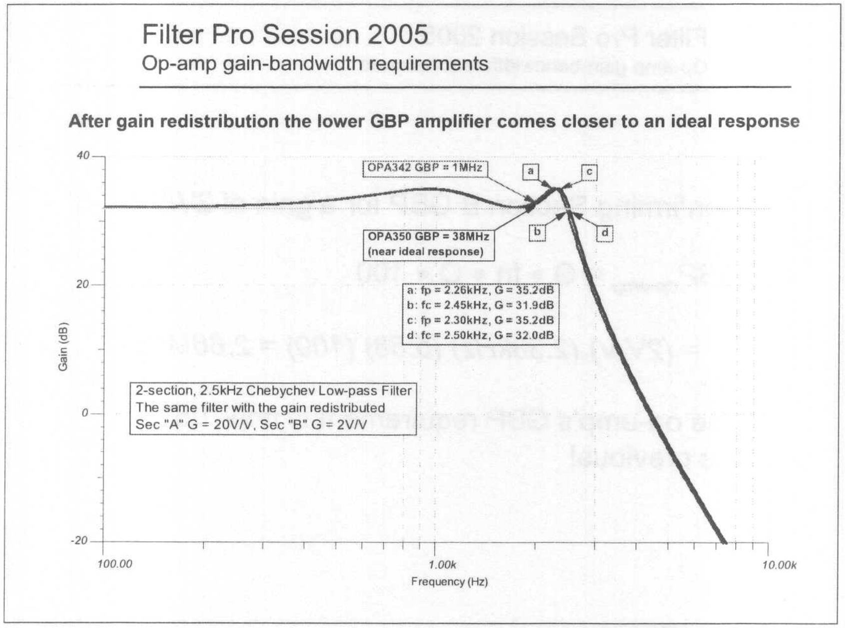
Confirming Section B GBP for a gain of 2V/V:

$$GBP_{op\text{-}amp} = G \bullet fn \bullet Q \bullet 100$$

$$= (2V/V)\ (2.38kHz)\ (5.58)\ (100) = 2.66MHz$$

The op-amp's GBP requirement is now 1/5th the previous!

With a near optimal redistribution of gains the required op-amp GBP is now one-fifth the previous value.



Filter Pro Session 2005

Op-amp gain-bandwidth requirements

After gain redistribution the lower GBP amplifier comes closer to an ideal response

OPA342 GBP = 1MHz

OPA350 GBP = 38MHz
(near ideal response)

a: fp = 2.26kHz, G = 35.2dB
b: fc = 2.45kHz, G = 31.9dB
c: fp = 2.30kHz, G = 35.2dB
d: fc = 2.50kHz, G = 32.0dB

2-section, 2.5kHz Chebychev Low-pass Filter
The same filter with the gain redistributed
Sec "A" G = 20V/V, Sec "B" G = 2V/V

The same two op-amps are used in this filter simulation, but with the gain optimally redistributed. The OPA350 response is unchanged and still near ideal, while the OPA342 with its much lower bandwidth comes very close to providing an ideal response. The gain is only off by about 0.1dB and the frequency inaccuracy is only about 2%.

The OPA340 or OPA341 op-amps with 5.5MHz GBP would serve well in this filter.

Filter Pro Session 2005

Where to obtain Filter Pro:

http://focus.ti.com/docs/toolsw/folders/print/filterpro.html

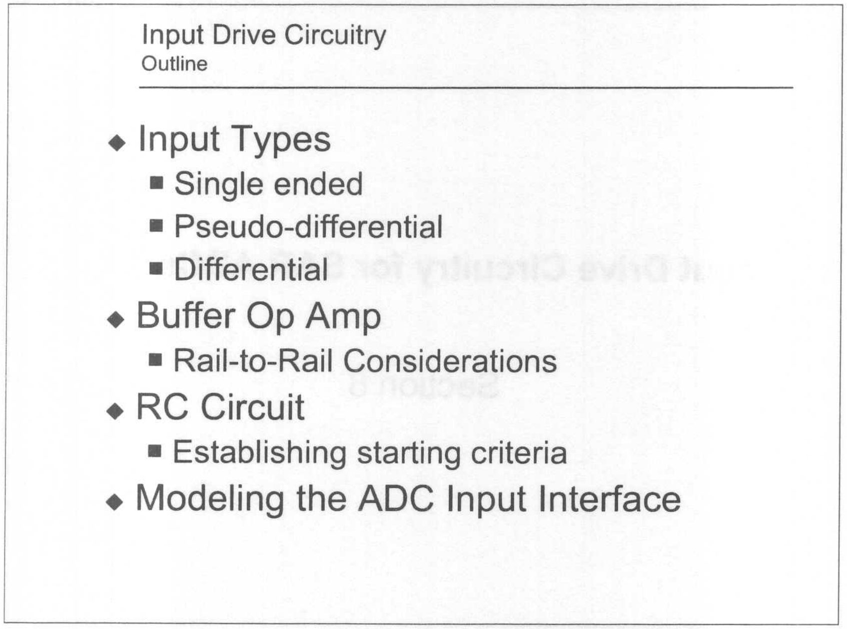
or go to: www.ti.com

and enter Filter Pro in the keyword search box

Input Drive Circuitry for SAR ADCs

Section 8

SAR ADCs in particular have input stages that have a very dynamic behavior. Designing circuitry to drive these loads is an interesting challenge. We've been looking at this for some time now, and are making progress toward a better understanding of all of the factors at play in the op amp to ADC interface.



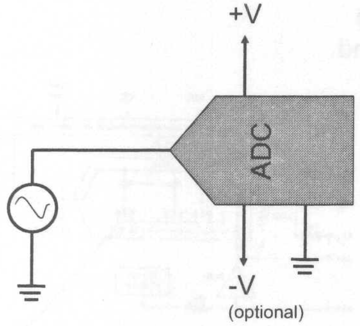
Input Drive Circuitry

Outline

- Input Types
  - Single ended
  - Pseudo-differential
  - Differential
- Buffer Op Amp
  - Rail-to-Rail Considerations
- RC Circuit
  - Establishing starting criteria
- Modeling the ADC Input Interface

Input Drive Circuitry

Input Stages: Single-Ended Inputs



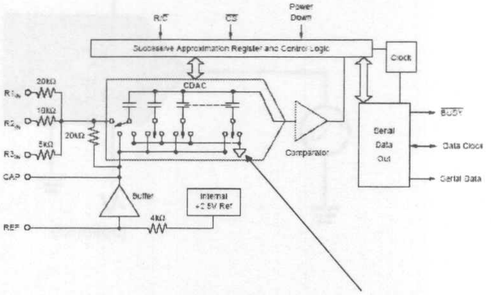
Before we look at the op amp closely, its important to know what type of ADC input you have – many options are available.

Shown above is a single-ended input. These inputs reference the input signal to ground. Depending upon the converter, the actual input voltage may be allowed to swing above and below the converter's ground reference, or may be required to stay only above ground.

Input Drive Circuitry

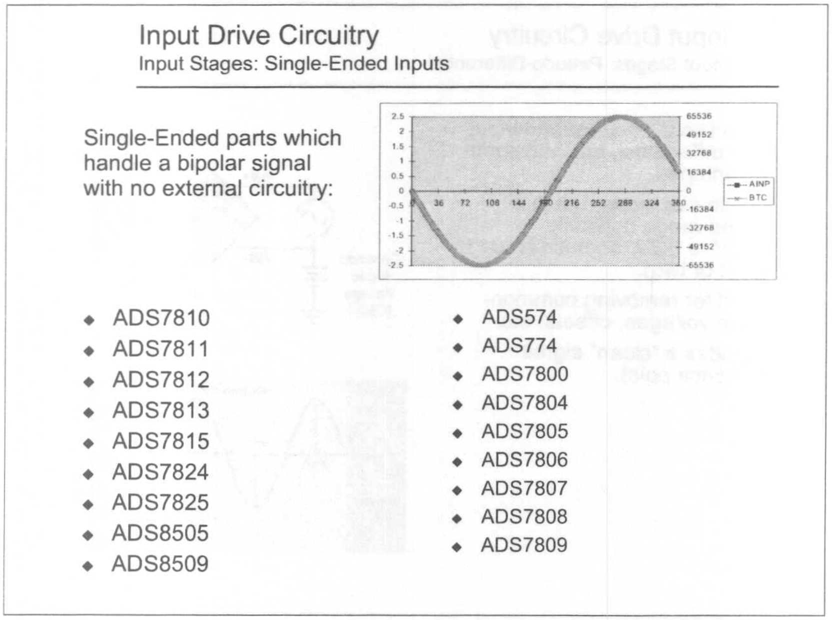
Input Stages: Single-Ended Inputs

- A single input pin for a particular channel.
- Input signals are referred to ground.



Note: input is referred to GROUND!

Single-ended inputs are easy to spot: there's only one input pin for an input signal! The other reference point is the converter's ground pin.


2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
0
36
72
108
144
180
216
252
288
324
360
65536
49152
32768
16384
0
-16384
-32768
-49152
-65536
AINP
BTC

Input Drive Circuitry

Input Stages: Single-Ended Inputs

Single-Ended parts which handle a bipolar signal with no external circuitry:

- ADS7810
- ADS7811
- ADS7812
- ADS7813
- ADS7815
- ADS7824
- ADS7825
- ADS8505
- ADS8509

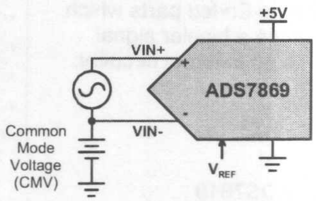
- ADS574
- ADS774
- ADS7800
- ADS7804
- ADS7805
- ADS7806
- ADS7807
- ADS7808
- ADS7809

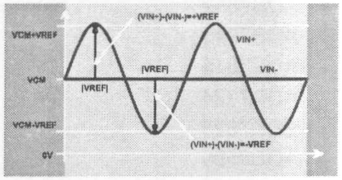
Parts which handle a signal that moves both above and below ground are said to have a *bipolar* input; the input signal may take one of two polarities, above and below ground. Texas Instruments offers a broad line of bipolar input converters.

Input Drive Circuitry

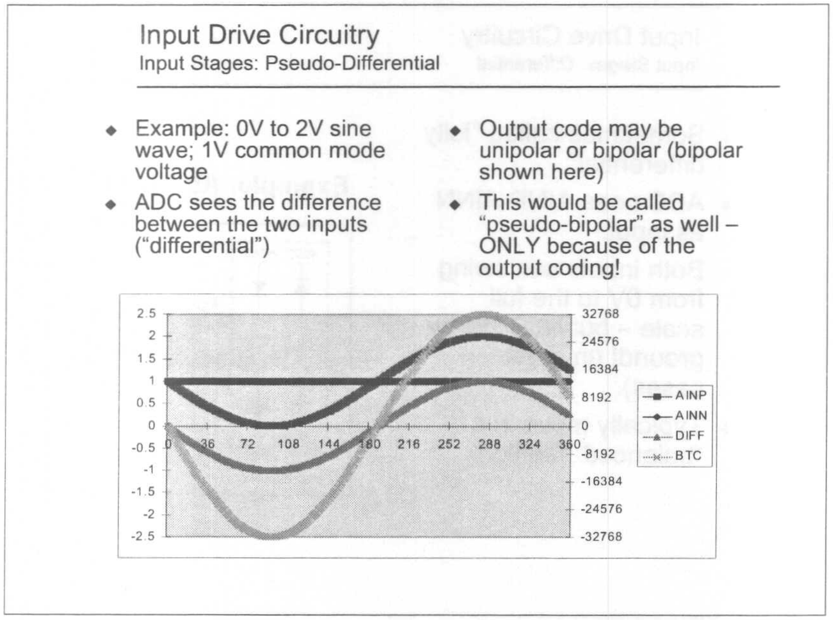
Input Stages: Pseudo-Differential

- More like a single-ended input than differential, but with some advantages.
- IN- pin can move, but in a limited range (typically -0.2V to 0.2V, sometimes up to a volt or two).
- Good for removing common-mode voltages, offsets, etc.
- Provides a “clean” signal reference point.





Some of our converters offer what is called a *pseudo-differential* input. In this type of input, the signal is referenced to a second input pin. This input pin can only accept a small range of voltages, perhaps a few hundred millivolts. This can be very helpful in situations where the signal has a slight common-mode offset, as this offset gets removed because the converter sees only the difference between the positive input pin and the negative input pin.



Input Drive Circuitry

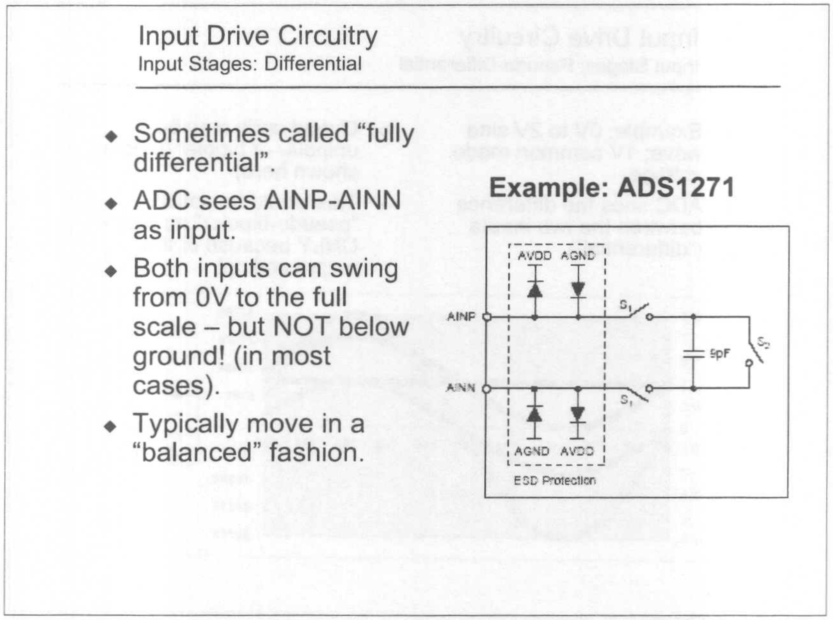
Input Stages: Pseudo-Differential

- Example: 0V to 2V sine wave; 1V common mode voltage
- ADC sees the difference between the two inputs (“differential”)
- Output code may be unipolar or bipolar (bipolar shown here)
- This would be called “pseudo-bipolar” as well – ONLY because of the output coding!

2.5
2
1.5
1
0.5
0
-0.5
-1
-1.5
-2
-2.5
0
36
72
108
144
180
216
252
288
324
360
32768
24576
16384
8192
0
-8192
-16384
-24576
-32768
AINP
AINN
DIFF
BTC

Here’s an example of a pseudo-differential input at work: a sine wave that swings from 0V to 2V is connected to the positive input, while the 1V common-mode signal is connected to the negative input. The ADC only sees the difference between the two pins, so it effectively sees a ±1V input. With an output coding that takes on positive and negative values, the unipolar 0V to 2V signal could be interpreted by the system as the ±1V sine wave - without needing bipolar power supplies!

slide could be
wrong



Input Drive Circuitry

Input Stages: Differential

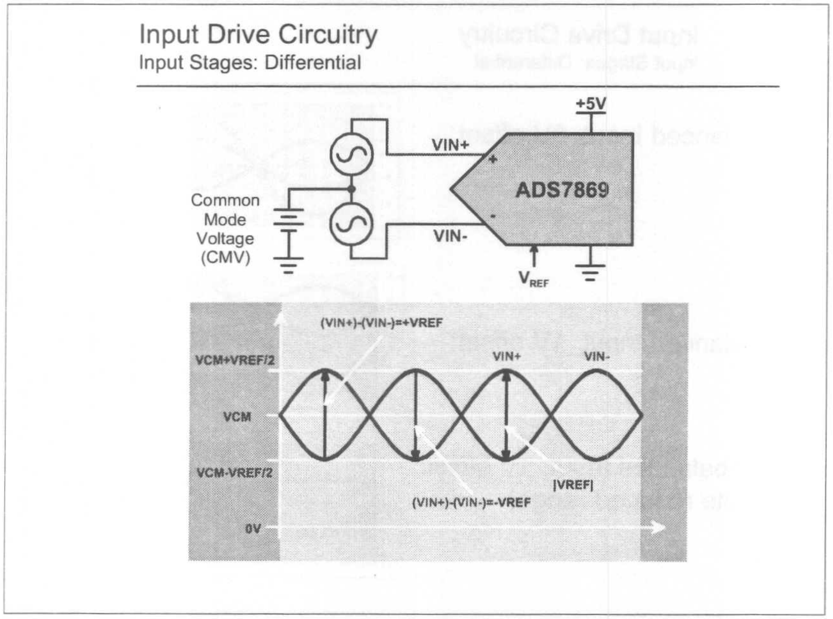
- Sometimes called “fully differential”.
- ADC sees AINP-AINN as input.
- Both inputs can swing from 0V to the full scale – but NOT below ground! (in most cases).
- Typically move in a “balanced” fashion.

Example: ADS1271

AVDD
AGND
AINP
S1
5pF
S2
AINN
S1
AGND
AVDD
ESD Protection

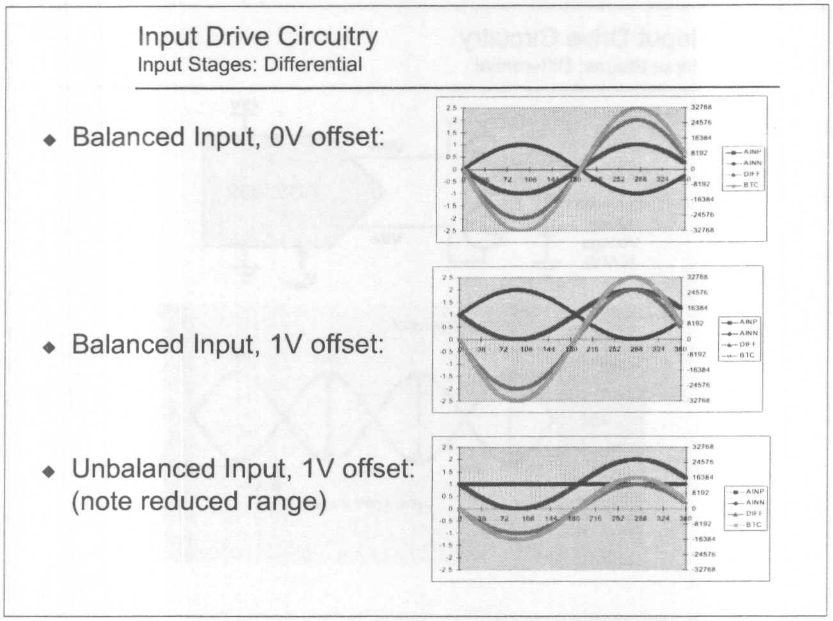
While the previous input circuit was a differential input, since it took the difference signal between the two input pins, it was considered “pseudo-differential” because one input was limited in range. In a truly differential input stage, both input pins can swing the full range, and typically move in a balanced fashion – as one input goes up, the other goes down in a corresponding way.

These types of inputs are commonly found on single-supply converters, such as delta-sigma or pipeline converters. The differential input offers the advantages of common-mode rejection, as well as limiting the input voltage swing required on each pin while preserving a high dynamic range.



Input Drive Circuitry
Input Stages: Differential

This slide illustrates the balanced fashion that differential inputs are typically used with.



Input Drive Circuitry

Input Stages: Differential

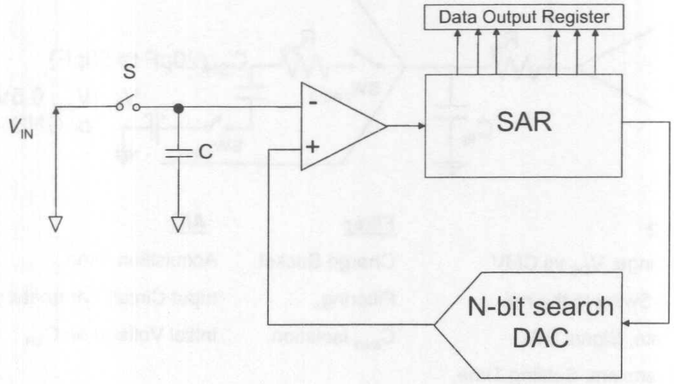
- Balanced Input, 0V offset:
- Balanced Input, 1V offset:
- Unbalanced Input, 1V offset: (note reduced range)

This slide shows the power of the differential input: with or without an input common-mode voltage, the converter sees the difference of the two pins. Each pin needs to swing only 1V, yet the converter sees a 2V swing in each direction.

However, it is important to recognize that using such a converter in a single-ended mode, by grounding or fixing one input at a fixed voltage, (not using it in a balanced fashion) results in a loss of dynamic range.

Input Drive Circuitry

Successive Approximation Converters



The most popular and versatile converter is the Successive Approximation Register (SAR) type. These converters work by comparing the analog signal voltage to known fractions of the full scale voltage and setting or clearing bits in a data register as a consequence.

Modern SARs use a C-DAC to successively compare bit combinations, set or clear the corresponding bits in a data register, and they also tend to have an integrated sample/hold function.

A typical SAR conversion cycle has two phases; a sampling phase and a conversion phase.

During the sampling phase, the analog input signal is allowed to charge the ADC's Sample-and-Hold (S/H) capacitor to a level proportional to the analog input.

Conversion begins immediately following the sampling phase. Conversion successively compares the unknown value of the charge stored in the S/H capacitor to known fractions of charge. After each comparison, logic on the ADC determines if the unknown charge is greater or smaller than the known fractional charge. The process will be like this

x >1/2 FS ? - Y - set the corresponding bit

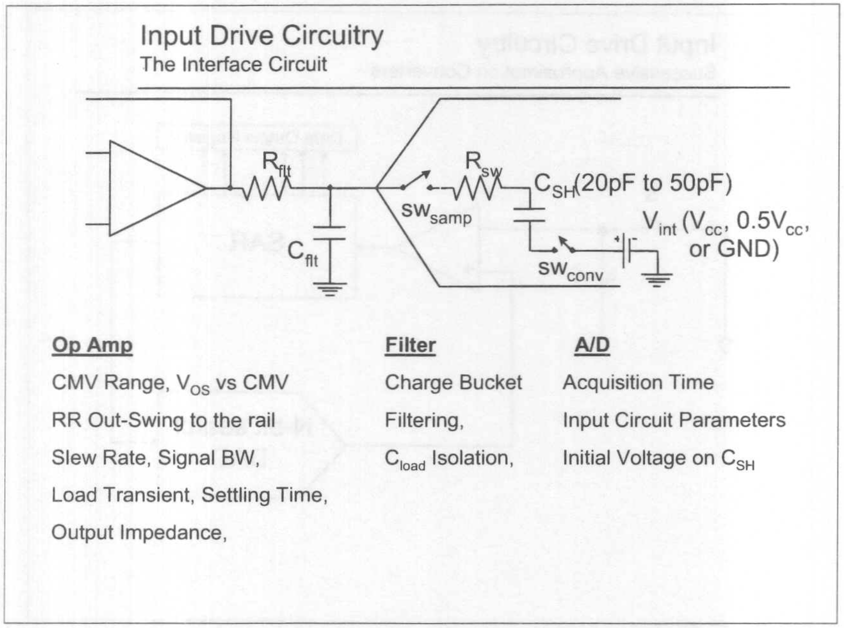
x >3/4 FS ? - N - clear the corresponding bit

x >5/8 FS ? - Y - set the corresponding bit

etc

At the end of the process the data register will contain a binary value proportional to the value initially placed on the S/H capacitor. The user reads this value out as converted data.

As shown in the diagram above, these converters rarely provide any type of built-in input buffer amplifier. So choosing an amplifier to match up to these converters is what we will examine next.



Input Drive Circuitry

The Interface Circuit

Op Amp	**Filter**	**A/D**
CMV Range, V_{OS} vs CMV	Charge Bucket	Acquisition Time
RR Out-Swing to the rail	Filtering,	Input Circuit Parameters
Slew Rate, Signal BW,	C_{load} Isolation,	Initial Voltage on C_{SH}
Load Transient, Settling Time,		
Output Impedance,		

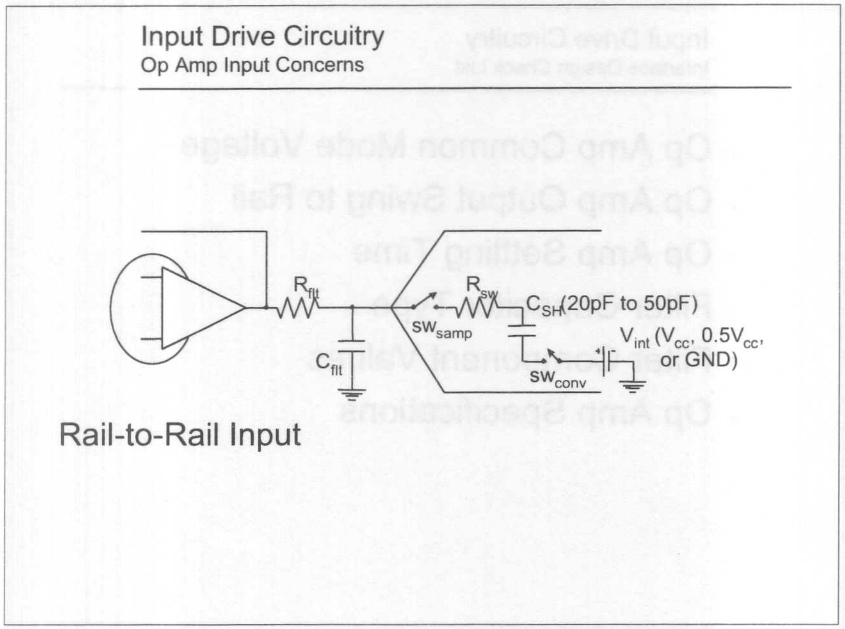
A closer look at the SAR ADC input and drive circuit is shown here. Many considerations come into play in choosing the optimum external R and C values, and choosing the best op amp to drive this circuit. We'll examine these in more detail in the following slides.

Input Drive Circuitry

Interface Design Check List

- Op Amp Common Mode Voltage
- Op Amp Output Swing to Rail
- Op Amp Settling Time
- Filter Capacitor Type
- Filter Component Values
- Op Amp Specifications

Some of the main factors we need to concentrate on are listed here.



Input Drive Circuitry

Op Amp Input Concerns

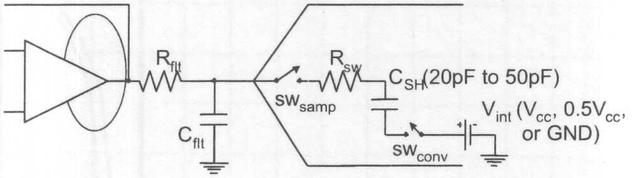
Rail-to-Rail Input

At the op amp input, one of the first considerations, especially with single-supply op amps, is how close to the power supply rail the input signals can swing. Note that many “rail-to-rail” input op amps really mean that the input can swing only within a few hundred millivolts of the rail.

Beware too of rail-to-rail op amps which use two different input stages to handle the rail-to-rail swing. See the Analog e-Lab (September 2004:” Avoiding the Pitfalls with Single Supply Op Amps”, http://www.eetimes.com/netseminar.html) to see how using these devices in a noninverting configuration can lead to significant distortion and errors. If faced with using one of these op amps, consider using them only in an inverting amplifier configuration.

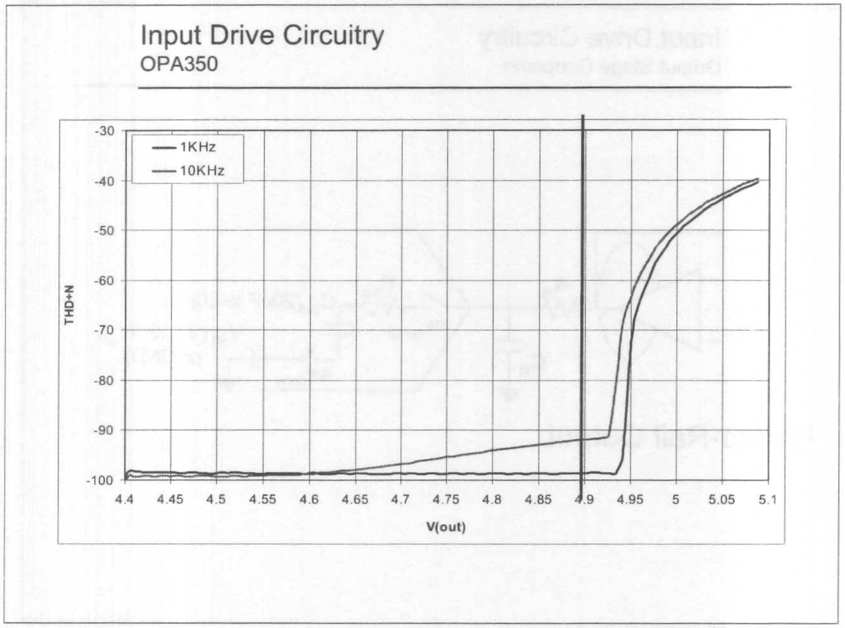
Input Drive Circuitry

Output Stage Concerns



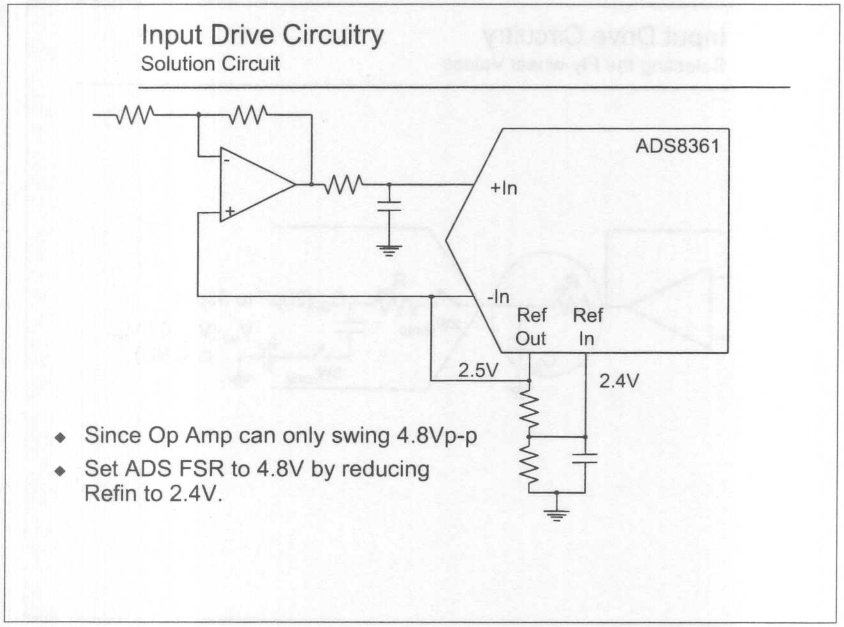
Rail-to-Rail Output

Similarly, the output stage of a "rail-to-rail" amplifier may not actually swing all the way to the supply rail. This can impact the dynamic range that can be used by the SAR ADC, many of which feature a true rail-to-rail input.



Input Drive Circuitry
OPA350

The plot above shows distortion versus output voltage swing of an OPA350, a popular rail-to-rail op amp. With a +5V power supply, you can see that noticeable distortion begins to appear when you get within 400mV of the rail. This is in spite of the fact that the OPA350 is rated to swing within 100mV of the rail (red line on this graph)! While the output can indeed swing that close to the rail, the signal fidelity there may not be optimal.

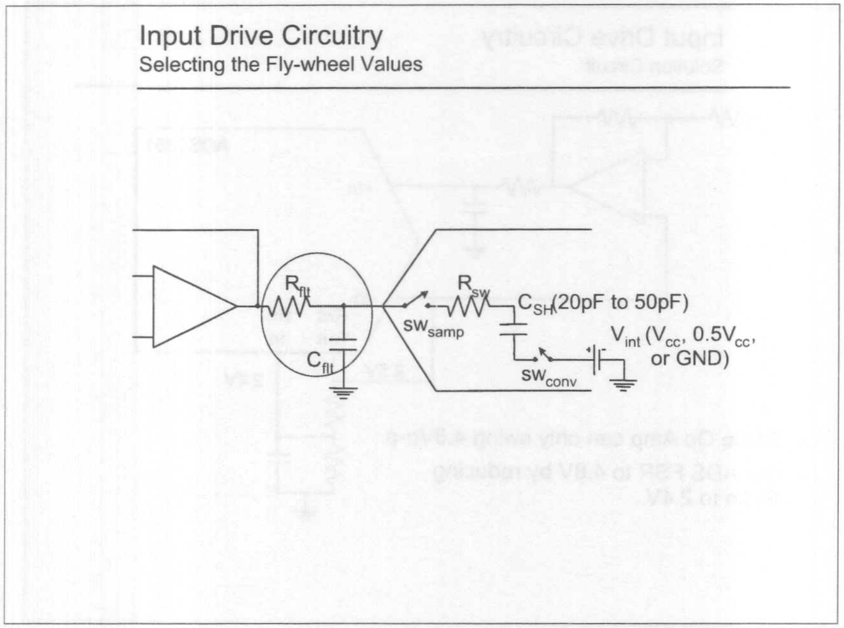


Input Drive Circuitry

Solution Circuit

- Since Op Amp can only swing 4.8Vp-p
- Set ADS FSR to 4.8V by reducing Refin to 2.4V.

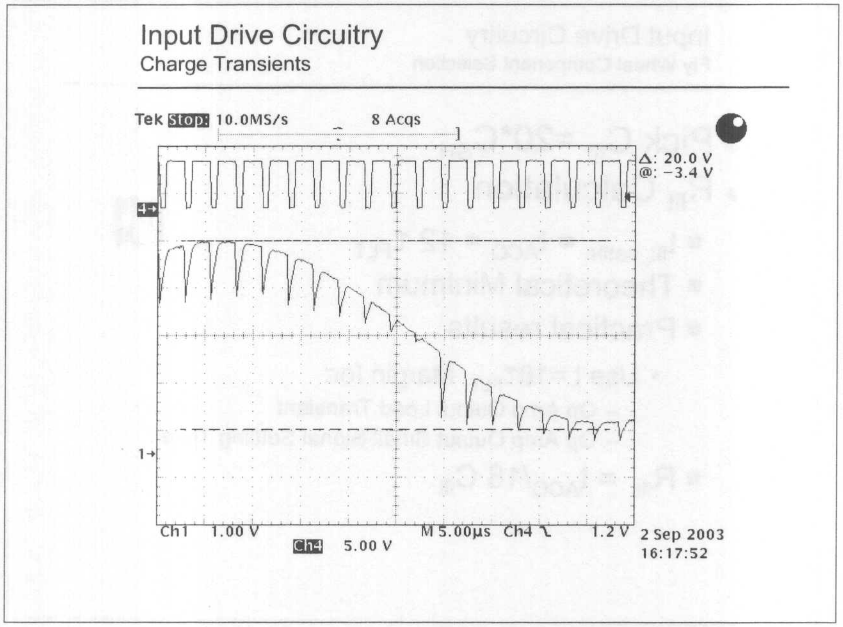
Here's a solution for dealing with the limitations of rail-to-rail op amps. To prevent the input stage from distorting the signal, the buffer amplifier is configured to be inverting, and the noninverting input is fixed at mid-rail (2.5V). However, because of the limitations in the output stage, the output of this op amp shouldn't swing more than ±2.4V around this 2.5V - a 4.8Vp-p swing. If we used the 2.5V reference as the reference for the ADC, then we would be giving up 200mV of dynamic range! By setting the ADC reference to 2.4V, the ADC then has a full-scale range of 4.8V, exactly matching what our op amp is capable of - and preserving dynamic range.



Input Drive Circuitry

Selecting the Fly-wheel Values

Now that we've handled some of the issues we're concerned with about the driving op amp, we need to examine the RC circuit we use in front of the SAR ADC – what's come to be called the "flywheel" circuit. It is so named because one of the functions of this circuit is for it to act as a charge-storage reservoir-storing energy, like a flywheel in a mechanical system, to charge the internal sampling capacitor of the SAR input stage.



Input Drive Circuitry

Charge Transients

The input sampling capacitor of a SAR ADC is connected to the input signal during the sampling or acquisition phase. The input signal is captured on this capacitor, then the sampling switch opens, disconnecting the input from the rest of the ADC. The conversion process then takes place, acting on this stored voltage, usually by redistributing charge from the capacitor to other capacitors in the internal DAC made up of capacitors.

When the ADC is ready to take another sample, the sampling switch closes again, reconnecting the input to the sampling capacitor. Any residual charge on this capacitor will have only one place to go – back out the input pin of the ADC!

The scope photo above shows these charge injection transients. This was generated by placing a 10KΩ resistor between the buffer op amp and the SAR ADC input, so we could clearly see the spikes. In general, the input impedance presented to a SAR converter should never be that high. Even with lower impedances, you may see spikes at the input of your SAR ADC. The spikes themselves are not bad for the converter, but if they do not settle back to the correct input value within the acquisition time of the converter, then this may result in a measurement error. These spikes also present a very demanding load to the driving op amp.

This is the second function of the RC filter we place in front of the SAR – to provide a path for this charge injection to go, and to do some minimal isolation of the op amp output to these transients.

Input Drive Circuitry

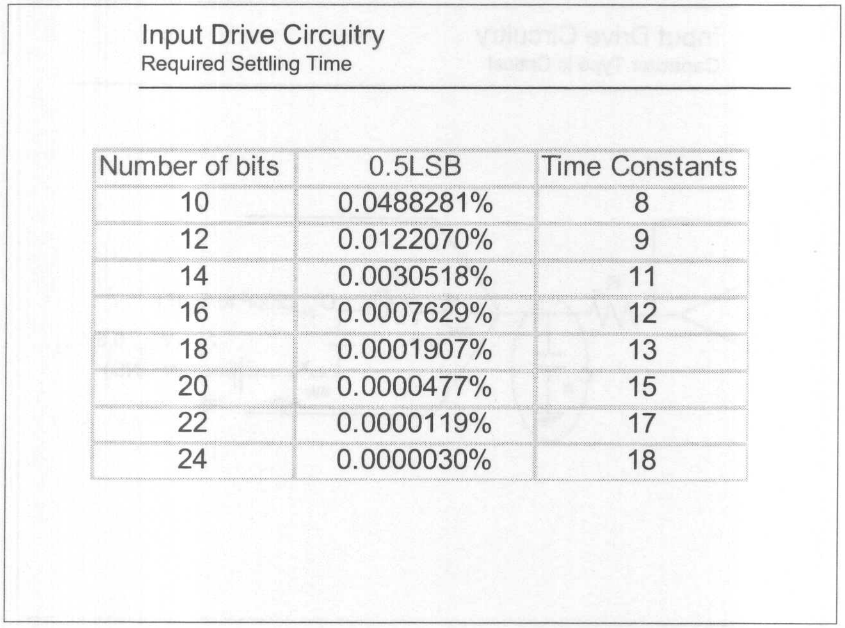
Fly Wheel Component Selection

- Pick $C_{flt} = 20 * C_{SH}$
- R_{flt} Calculation
  - $t_{flt\_settle} = t_{ACQ} = 12\ \tau_{FLT}$
  - Theoretical Minimum
  - Practical results
    - Use $t = 18\tau_{FLT}$ Margin for:
      - Op Amp Output Load Transient
      - Op Amp Output Small Signal Settling Time
  - $R_{flt} = t_{ACQ}/18\ C_{flt}$

In our 2004 seminar, we showed a detailed analysis of how to arrive at good starting values for the external RC components. A summary of the results is shown here.

To minimize the effects of nonlinearities in the internal sampling capacitor, as well as minimize the charge injection's effect on the input voltage, we choose the external capacitor value to be at minimum twenty times the size of the internal sampling capacitor.

Once the C value is chosen, the R value can be found by noting that this RC filter must fully settle to the desired accuracy within the acquisition time of the converter. For example, to achieve 16-bit settling accuracy, 12 time constants of that RC must be allowed. That is a theoretical minimum; practical experience tells us that you need even more time, so we add a 50% margin. Thus, for our example 16-bit case, 18 time constants must be able to go by within the acquisition time of our converter.



Input Drive Circuitry

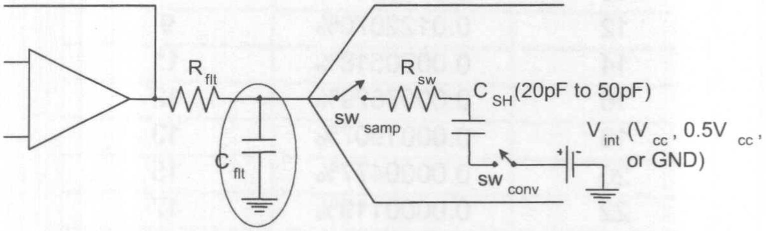
Required Settling Time

Number of bits	0.5LSB	Time Constants
10	0.0488281%	8
12	0.0122070%	9
14	0.0030518%	11
16	0.0007629%	12
18	0.0001907%	13
20	0.0000477%	15
22	0.0000119%	17
24	0.0000030%	18

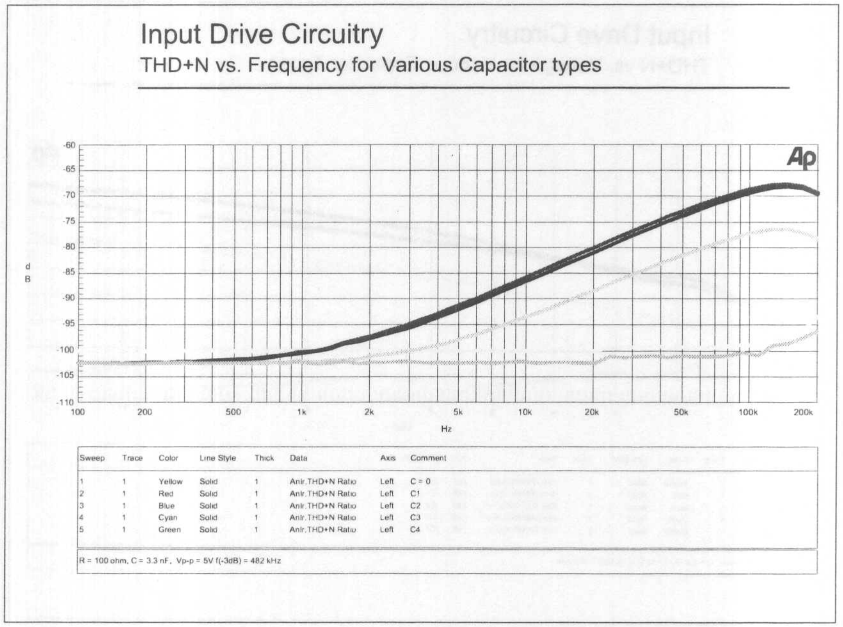
This chart details the number of time constants that must be allowed to achieve a certain accuracy in settling for different resolutions.

Input Drive Circuitry

Capacitor Type is Critical



Now that we've determined some starting values for the R and the C, let's consider what type of capacitor dielectric should be used for the external capacitor.



Input Drive Circuitry

THD+N vs. Frequency for Various Capacitor types

Sweep	Trace	Color	Line Style	Thick	Data	Axis	Comment
1	1	Yellow	Solid	1	Anlr.THD+N Ratio	Left	C = 0
2	1	Red	Solid	1	Anlr.THD+N Ratio	Left	C1
3	1	Blue	Solid	1	Anlr.THD+N Ratio	Left	C2
4	1	Cyan	Solid	1	Anlr.THD+N Ratio	Left	C3
5	1	Green	Solid	1	Anlr.THD+N Ratio	Left	C4

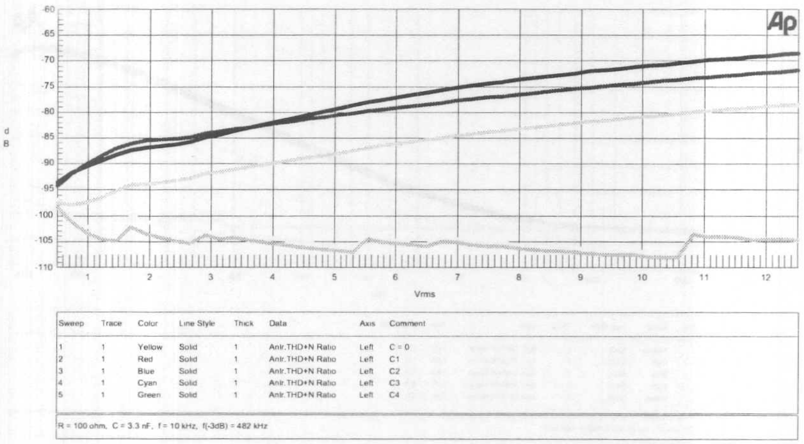
R = 100 ohm, C = 3.3 nF, Vp-p = 5V f(-3dB) = 482 kHz

Capacitors have a voltage coefficient, which means that the capacitance changes with applied voltage. This also tends to be nonlinear, and thus introduces distortion which changes with frequency. Here is a look at several capacitor types and their distortion versus frequency characteristics. The lowest line on this chart is the system measurement limit - and the line right next to that is a silver-mica capacitor. The other lines on this chart are from ceramic caps with different dielectrics — Z5U, Y5V, and X7R. Note that these types introduce significant distortion into the signal.

One ceramic dielectric, the C0G type, closely matches the silver mica performance (but is not shown on this chart).

Input Drive Circuitry

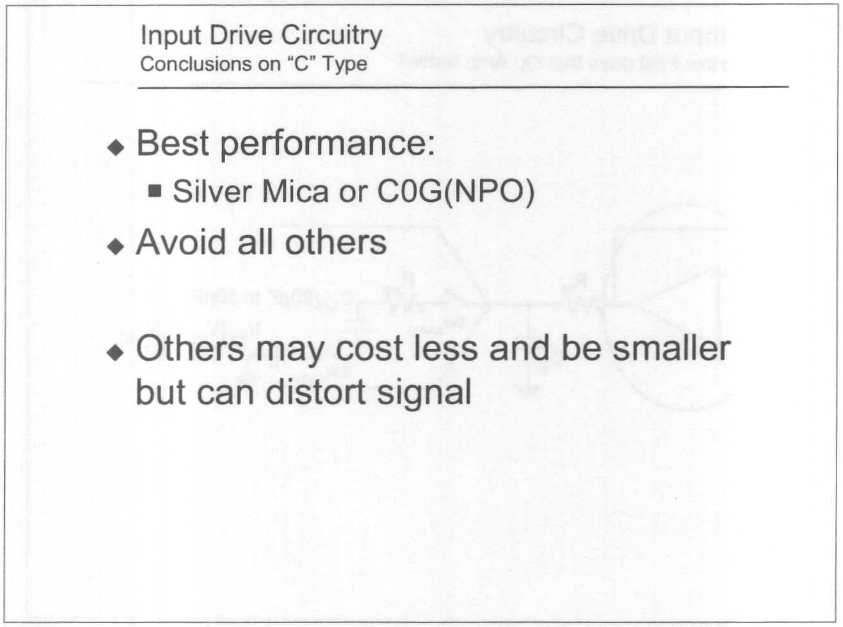
THD+N vs. Voltage for Various Capacitor Types



Sweep	Trace	Color	Line Style	Thick	Data	Axis	Comment
1	1	Yellow	Solid	1	Anlr.THD+N Ratio	Left	C = 0
2	1	Red	Solid	1	Anlr.THD+N Ratio	Left	C1
3	1	Blue	Solid	1	Anlr.THD+N Ratio	Left	C2
4	1	Cyan	Solid	1	Anlr.THD+N Ratio	Left	C3
5	1	Green	Solid	1	Anlr.THD+N Ratio	Left	C4

R = 100 ohm, C = 3.3 nF, f = 10 kHz, f(-3dB) = 482 kHz

This chart is similar to the previous one, except now we are looking at distortion versus voltage applied at a fixed frequency. Again, the sliver mica capacitor approaches the measurement limit of the test system.



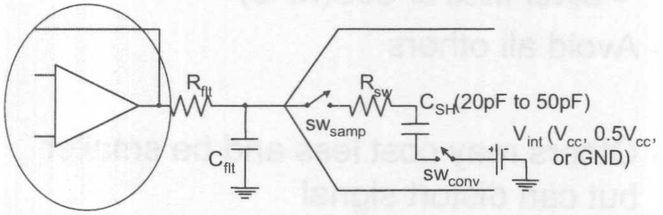
Input Drive Circuitry
Conclusions on “C” Type

- Best performance:
  - Silver Mica or C0G(NPO)
- Avoid all others
- Others may cost less and be smaller but can distort signal

From the previous data, we conclude that only silver mica or C0G ceramic capacitor types should be considered for use as the external capacitor. The other types may be less in cost but will hurt the overall system performance.

Input Drive Circuitry

How Fast does that Op Amp Settle?



We now return to the op amp, and consider some further issues with it, now that we've chosen the RC values. Of prime importance, obviously, is how quickly can the op amp settle? This is key as it must settle within the acquisition time of the converter.

Input Drive Circuitry

What Settling Time?

- Think of a linear voltage regulator, there are **TWO** Settling Times.
  - Line Transient
  - Load Transient
- **Same** applies here.
- Data sheets report settling time to 0.01% at best.

As we consider the op amp settling behavior, we must understand that an op amp has a settling time which is specified only for a change in the input signal, and how long the output takes to settle to within 0.01% - 12 bits. Higher resolution systems will need to allow more time.

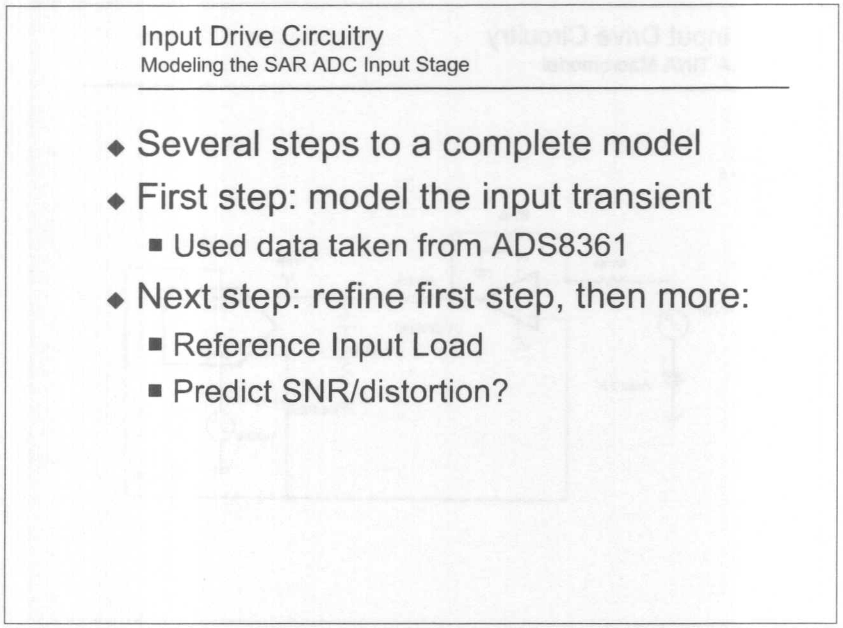
But this specified settling time is only for one behavior – we don't know anything about what happens to the op amp output as the load changes. Unlike a voltage regulator, where the response to both a line transient (input) and a load transient (output) is specified, we know only one of these for the op amp. Yet we have a circuit which we know clearly provides significant load transients to the op amp!

Input Drive Circuitry

Op Amp Specs to Check

- $UGBW_{mod} > 2^* 1/[2\pi R_{FLT} C_{FLT}]$
- Op Amp Transient Output Drive to R_{FLT} & C_{FLT}
  - $I_{Opk\ max} = (5\%\ V_{REF})/(R_{FLT})$

The Unity-Gain bandwidth (UGBW) of the op amp should be at least twice the bandwidth of the external RC filter. The op amp must have sufficient output drive current to charge the capacitor used.



Input Drive Circuitry

Modeling the SAR ADC Input Stage

- Several steps to a complete model
- First step: model the input transient
  - Used data taken from ADS8361
- Next step: refine first step, then more:
  - Reference Input Load
  - Predict SNR/distortion?

One thing that would be very nice to be able to do is to model this SAR drive circuitry in SPICE. This would allow for quick evaluations of different op amps and RC values to help choose the optimal values.

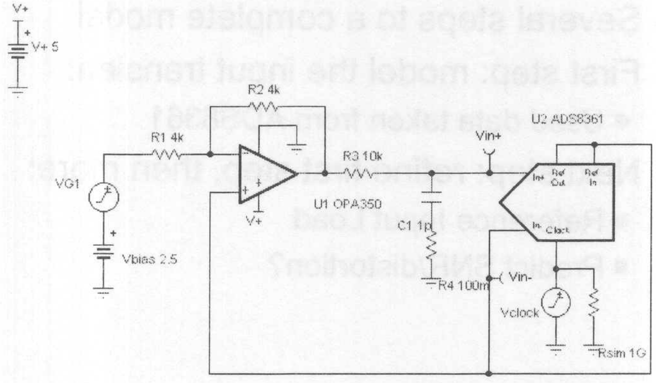
Of course, SPICE is only as good as the models you give it. Constructing such a model can be laborious - but we've made an attempt at modeling a SAR ADC input stage and created a macromodel we can use in TI-TINA.

The macromodel we've created uses data taken from our evaluations of the ADS8361, which has been extensively tested over the past few years as we've done this study. The model we have today is a preliminary look – we need to set up many different scenarios in TI-TINA, then verify them on the bench and compare the results to give us confidence that the model we have accurately reflects the behavior of the device. Refinements to the model can then be made to have it more closely reflect actual device performance.

Following this, we'd like to be able to also model the reference input load as this can be even more demanding than the input load on a SAR ADC. From these, it would be interesting to see if one could predict SNR or distortion using these models. These are all on our list of things to investigate.

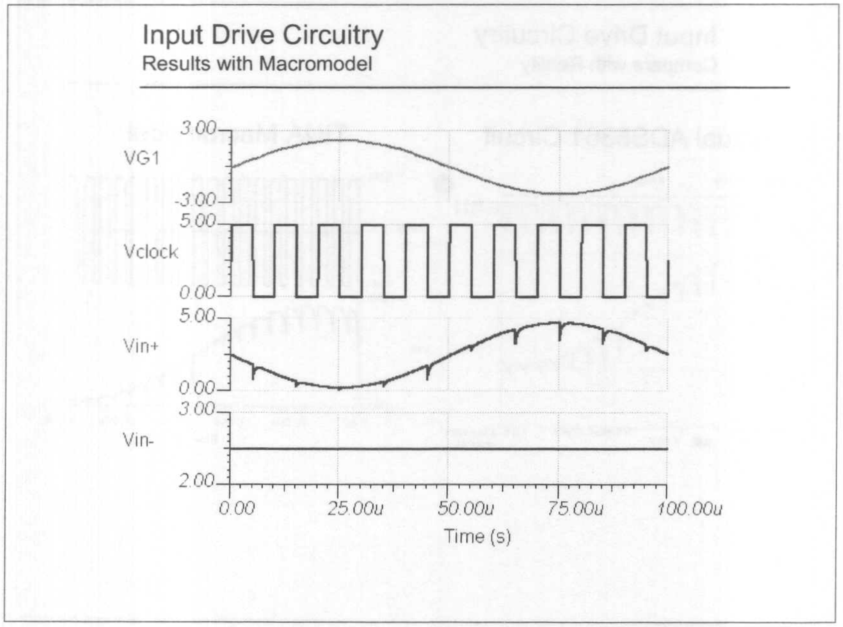
Input Drive Circuitry

A TINA Macromodel



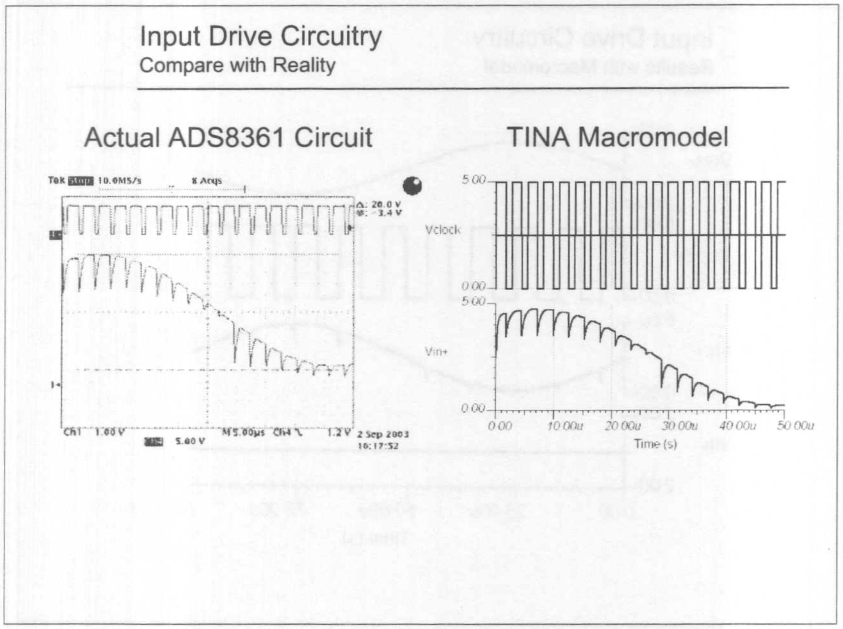
This is a look at a TINA circuit which uses this preliminary input stage model for the ADS8361. You'll notice that it is set up very much like the input stage design we showed earlier.

In this circuit, we've set the R to 10KΩ to mimic our first look at the charge transients, and set the C only to 1pF, in an attempt to model the scope probe capacitance used to take that data.



Input Drive Circuitry
Results with Macromodel

The results of a TINA simulation using this circuit clearly shows the load transients.



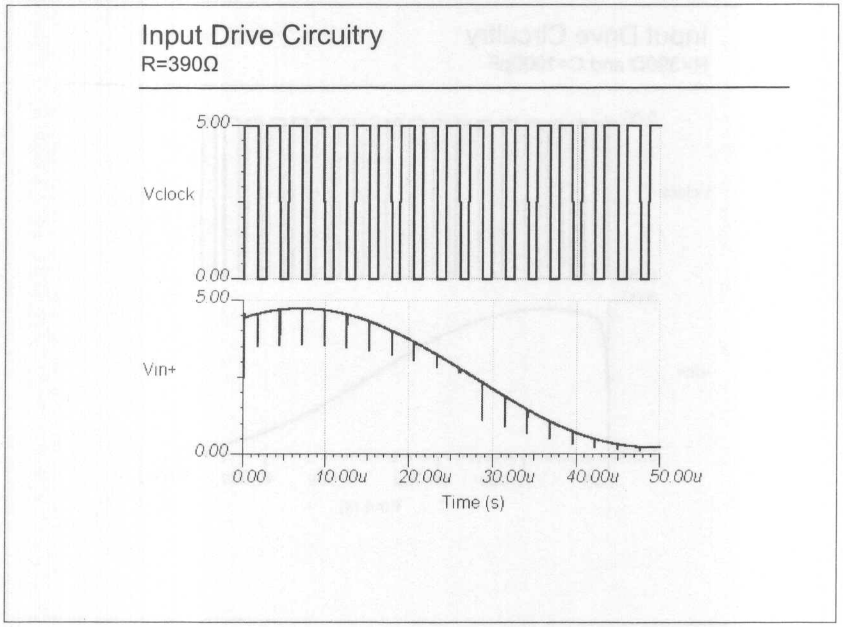
Input Drive Circuitry

Compare with Reality

Actual ADS8361 Circuit

TINA Macromodel

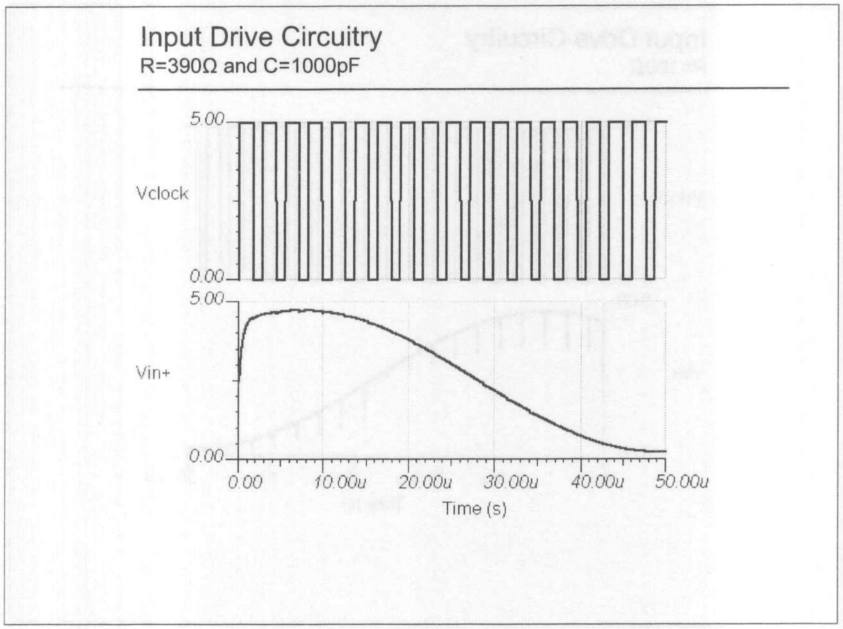
Comparing the actual circuit data with the simulation results from our TI-TINA model shows good agreement between what the model predicts and what we actually see in the lab.



Input Drive Circuitry

R=390Ω

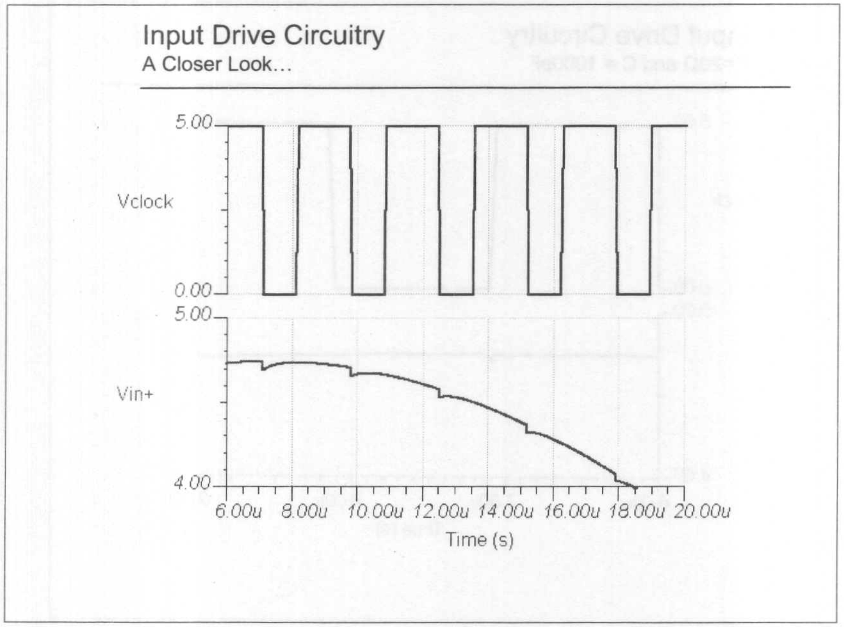
One question you might ask is, what happens if you don't have the "flywheel" capacitor, but bring the resistor down to a reasonable level? Here's a simulation using only 390Ω as the resistor, rather than 10KΩ. Notice that the spikes are still quite large, but it appears that the width of the spike has diminished.



Input Drive Circuitry

R=390Ω and C=1000pF

Now let's add the capacitor into the circuit – 1000pF with our 390Ω resistor. The input signal looks quite good now!

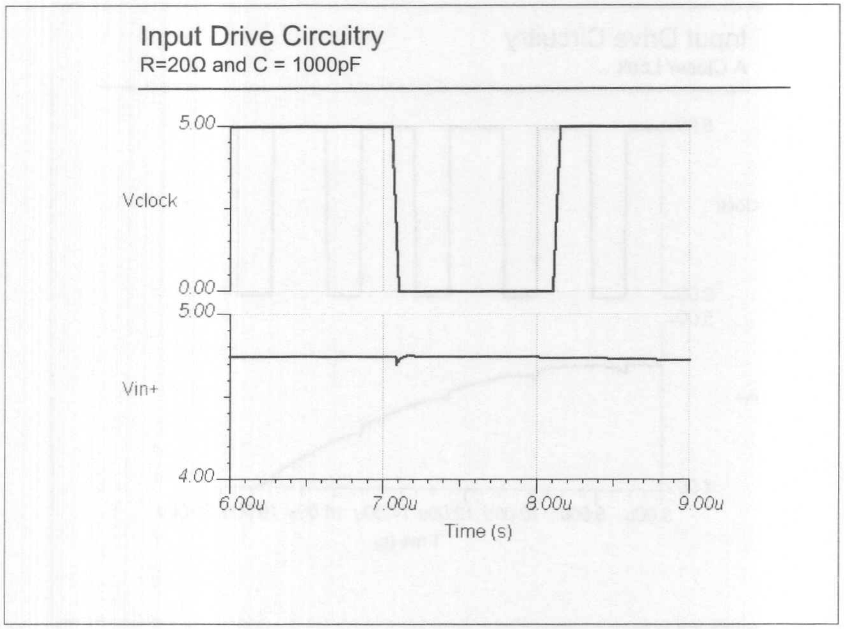


Input Drive Circuitry

A Closer Look…

If we zoom in on a section of the TI-TINA simulation results, we can see that there is still some effect on the input signal, even with the flywheel circuit in place. But remember, it's not so much the transient we need to remove, it is that we must be assured that it is fully settled before the acquisition time of the converter has passed.

This looks like it may be settled, but much higher resolution on our output results would be required to know if it's settled to 16 bits by the time the clock signal goes high again.



Input Drive Circuitry

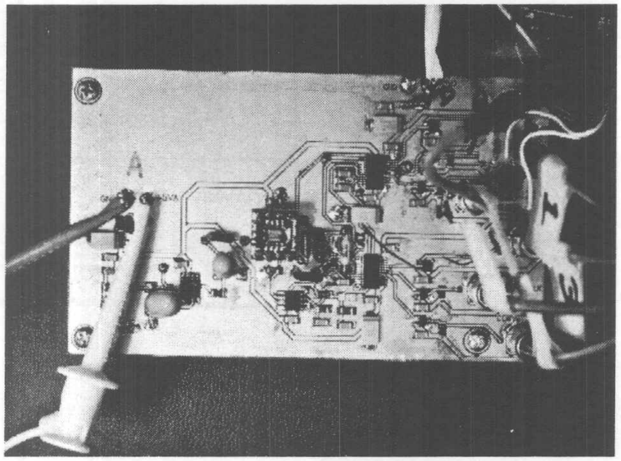
R=20Ω and C = 1000pF

In some cases, very low values of resistance are used. Here’s a look at the load transient with R dropped to 20 ohms. This looks like things have probably settled out well before the acquisition time is over.

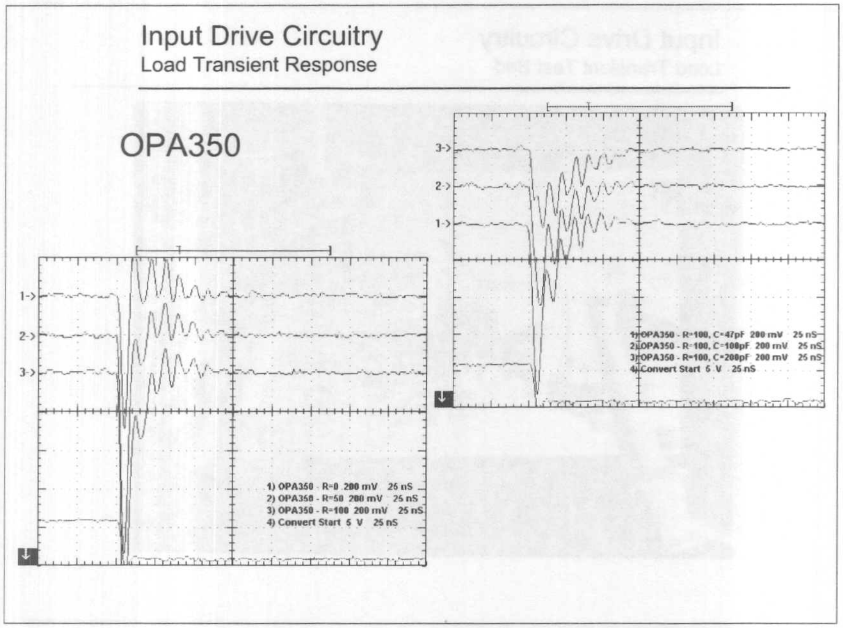
How can we verify this?

Input Drive Circuitry

Load Transient Test Bed



This is an experimental board that might allow us to actually measure the settling time of the input signal. It consists of two ADS8361s – one, the device used in the test, fed by an op amp and the RC filter. The second ADS8361 is looking at the same input. But the second ADS8361's timing is adjusted such that it converts slightly before or after the first converter. By stepping through conversion instants in 20ns steps, we can build a picture of the actual acquired values, and can see the settling time of the op amp while it is receiving the load transients from the main converter.

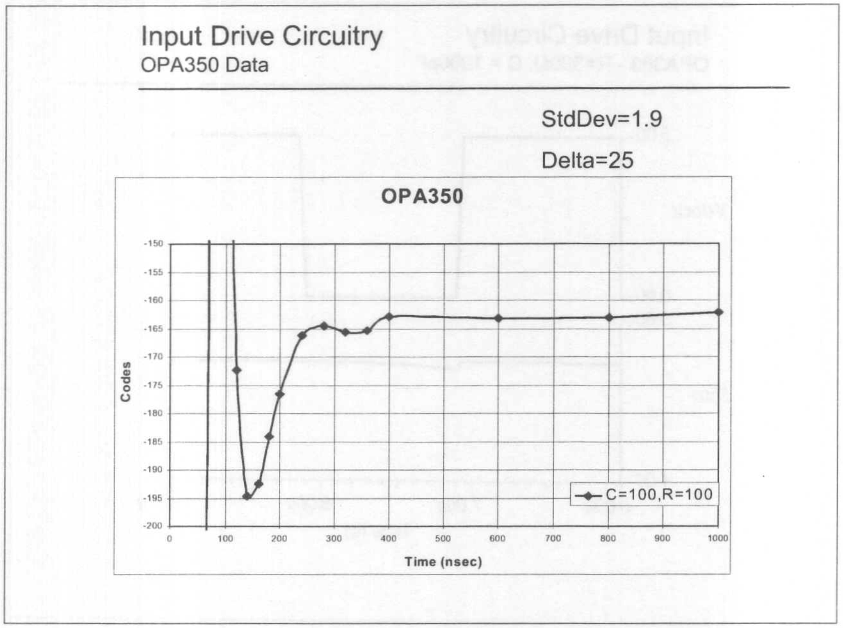


Input Drive Circuitry

Load Transient Response

OPA350

From looking at scope captures of the actual op amp under a transient load condition, we can see that there is significant ringing and settling. On the left, the settling with no external capacitor; on the right, different capacitors are used.



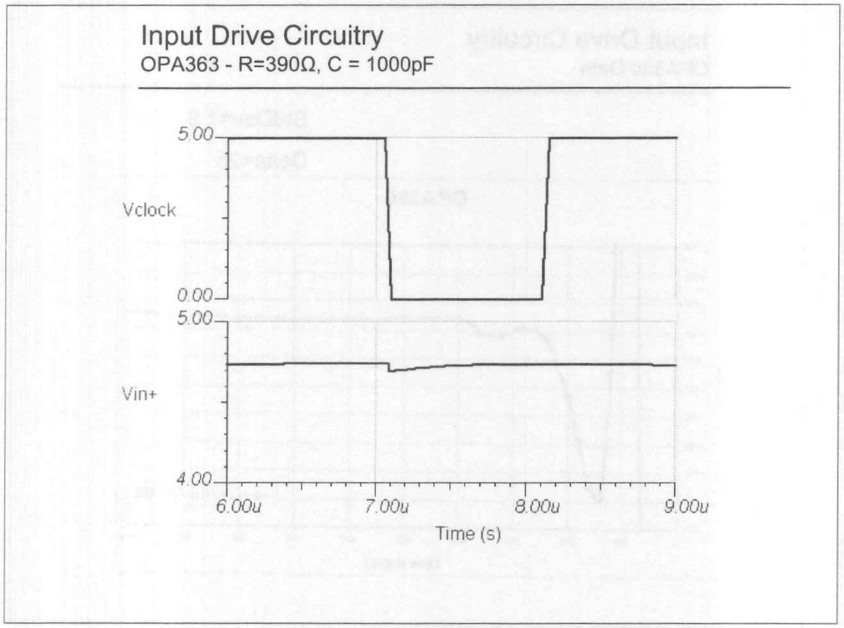
Input Drive Circuitry

OPA350 Data

StdDev=1.9

Delta=25

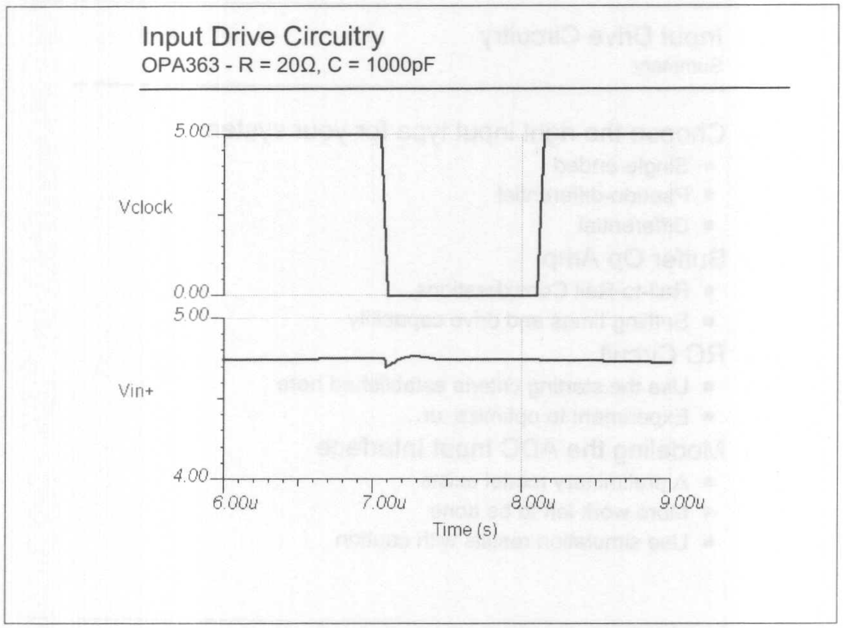
From our captured signals taken with the second ADC, we can “see” what the main ADC “sees” – a settling time of around 400ns to get within 2 codes of the final value, and longer to be truly “settled”. Note that the OPA350 macromodel used predicted much faster settling.



Input Drive Circuitry

OPA363 - R=390Ω, C = 1000pF

As with any modeling exercise, the results are only as good as the models used. The OPA350 macromodel does not model output characteristics very well; later models, such as the one for the OPA363, do a much better job of modeling output stage behavior. Here, we've used the OPA363 and can see the longer settling tail, although this is with a larger external R.

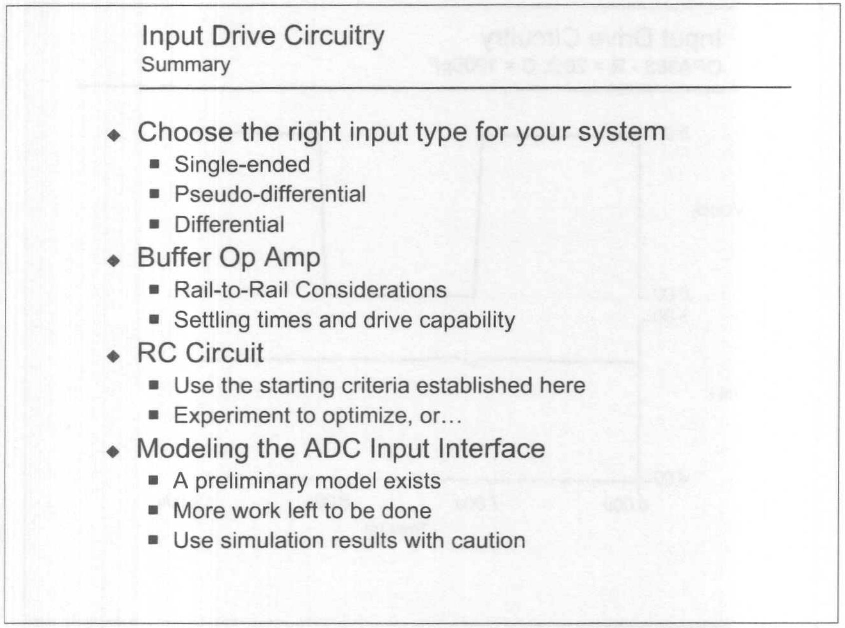


Input Drive Circuitry

OPA363 - R = 20Ω, C = 1000pF

Dropping the R value to the same condition as the previous model with the OPA350, we can see some of the ringing behavior beginning to appear. This is as a result of the better model available for the OPA363.

Much more development is necessary on both the op amp models and the ADC input model before these can be used as true predictors of what actually happens in a system. This preliminary look at things points out some of the areas where more work needs to be done.



Input Drive Circuitry

Summary

- Choose the right input type for your system
  - Single-ended
  - Pseudo-differential
  - Differential
- Buffer Op Amp
  - Rail-to-Rail Considerations
  - Settling times and drive capability
- RC Circuit
  - Use the starting criteria established here
  - Experiment to optimize, or…
- Modeling the ADC Input Interface
  - A preliminary model exists
  - More work left to be done
  - Use simulation results with caution

PCB Layout Tips for High Resolution

Section 9

Layout of printed circuit boards (PCBs) could be an all-day discussion. Rather than give you a set of rules to follow, which almost always have exceptions, we'll look here at some general principles and then review a number of actual board layouts that illustrate common mistakes.

PCB Layout Tips

Outline

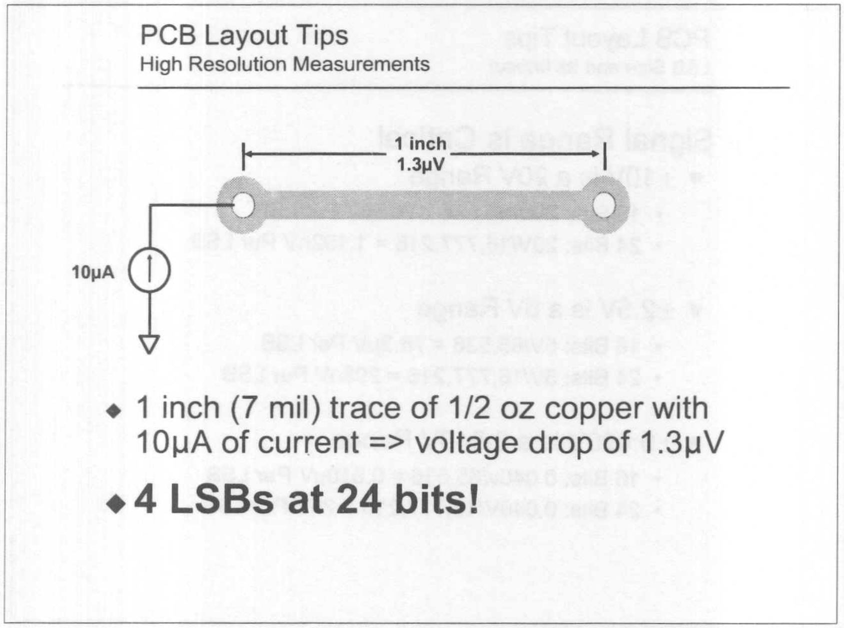
- LSB size
  - Impact on layout decisions
- Basic PCB Design Principles
  - For High Resolution
- PCB Layout Reviews
  - Learn from the misfortune of others!

PCB Layout Tips

LSB Size and Its Impact

- Signal Range Is Critical
  - ±10V is a 20V Range
    - 16 Bits: 20V/65,536 = 305µV Per LSB
    - 24 Bits: 20V/16,777,216 = 1,192nV Per LSB
  - ±2.5V is a 5V Range
    - 16 Bits: 5V/65,536 = 76.3µV Per LSB
    - 24 Bits: 5V/16,777,216 = 298nV Per LSB
  - ±0.020V is a 0.040V Range
    - 16 Bits: 0.040v/65,536 = 0.610µV Per LSB
    - 24 Bits: 0.040V/16,777,216 = 2nV Per LSB

Where ±10V gives you an LSB of 305µV, a 5V signal only gives you 76.2µV for a 16-bit LSB. When gain is added into the system, the resolution can sink into the noise.

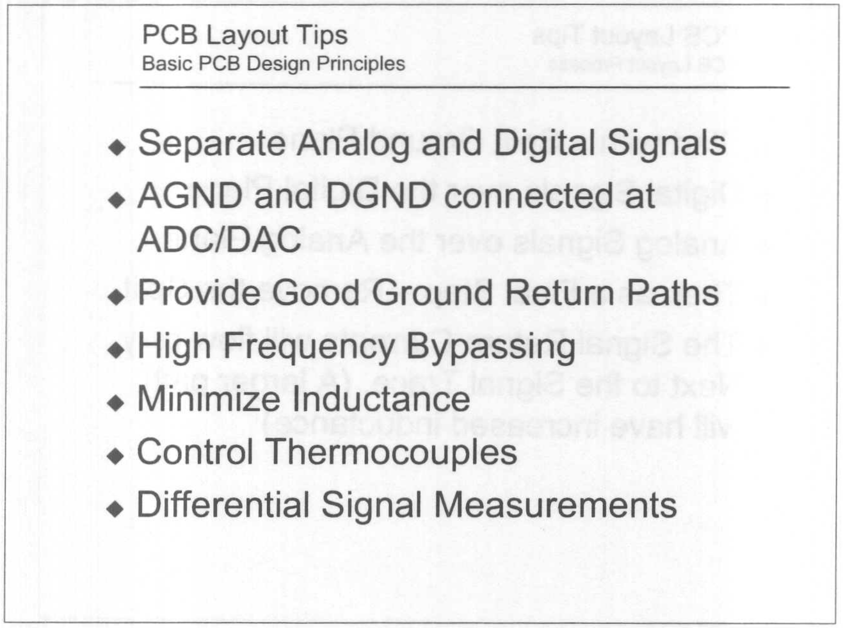


PCB Layout Tips

High Resolution Measurements

- 1 inch (7 mil) trace of 1/2 oz copper with 10µA of current => voltage drop of 1.3µV
- **4 LSBs at 24 bits!**

You must be careful with the PCB layout for a 12-bit design. For a 16-bit design you have to do everything correctly to achieve that performance; it should be obvious that a 24-bit design will be affected by any errors in the layout.



PCB Layout Tips

Basic PCB Design Principles

- Separate Analog and Digital Signals
- AGND and DGND connected at ADC/DAC
- Provide Good Ground Return Paths
- High Frequency Bypassing
- Minimize Inductance
- Control Thermocouples
- Differential Signal Measurements

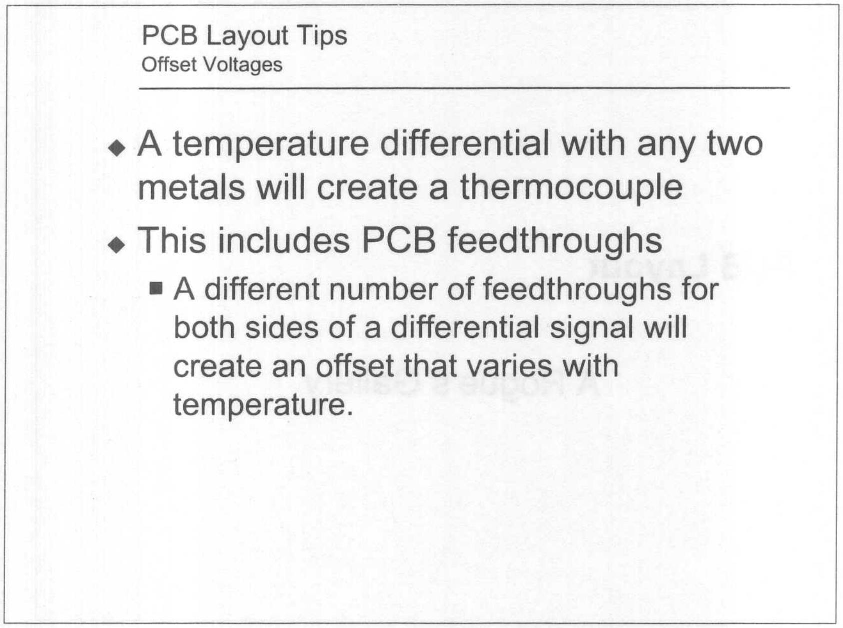
The separation of the Analog and Digital section of the PCB keeps noisy digital from the low level analog. However, they must come together at the ADC or DAC. It is important that the Analog Ground and the Digital Ground be connected together at the ADC or DAC. This allows a quick return for the ground currents as the analog and digital portions of the device communicate.

Inductance is determined by the size of the loop of current. Providing a path for return currents next to the signal trace will reduce the inductance. That is the advantage of a solid ground plane. If a reasonably solid plane can't be achieved with two layers, then more layers should be used.

PCB Layout Tips
PCB Layout Process

- Start with a Split Ground Plane
- Digital Signals over the Digital Plane
- Analog Signals over the Analog Plane
- Then as a Final Step – Remove the Split
- The Signal Return Currents will flow only Next to the Signal Trace. (A larger path will have increased inductance)

Since AGND and DGND are connected at the ADC/DAC and also at the power supply, there is a ground loop created. By having a solid ground plane this can be eliminated. The analog signals will still just flow in the analog portions of the ground plane and the digital signals will only flow in the digital portion of the ground plane. The reason why they will not travel throughout the board, is because a larger path has a larger inductance. The signals will travel through the path of lowest impedance.

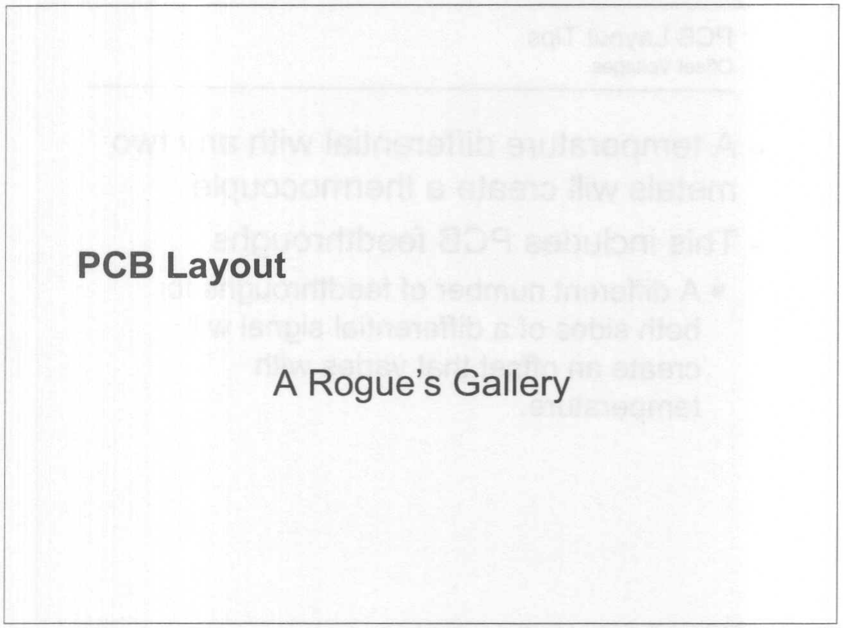


PCB Layout Tips

Offset Voltages

- A temperature differential with any two metals will create a thermocouple
- This includes PCB feedthroughs
    - A different number of feedthroughs for both sides of a differential signal will create an offset that varies with temperature.

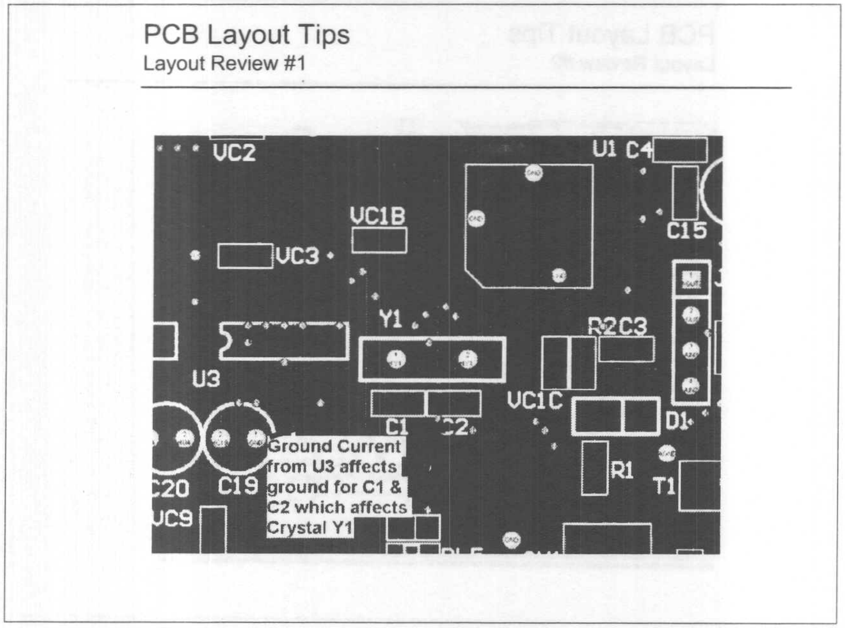
Your board design can create additional offset voltages because of the traces and feedthroughs. To minimize offset it is good to route the signal traces differentially next to each other so that they will see the same thermal gradients and the same number of feedthroughs.



PCB Layout

A Rogue's Gallery

We'll now look at a few examples of circuit layouts that make some common mistakes.

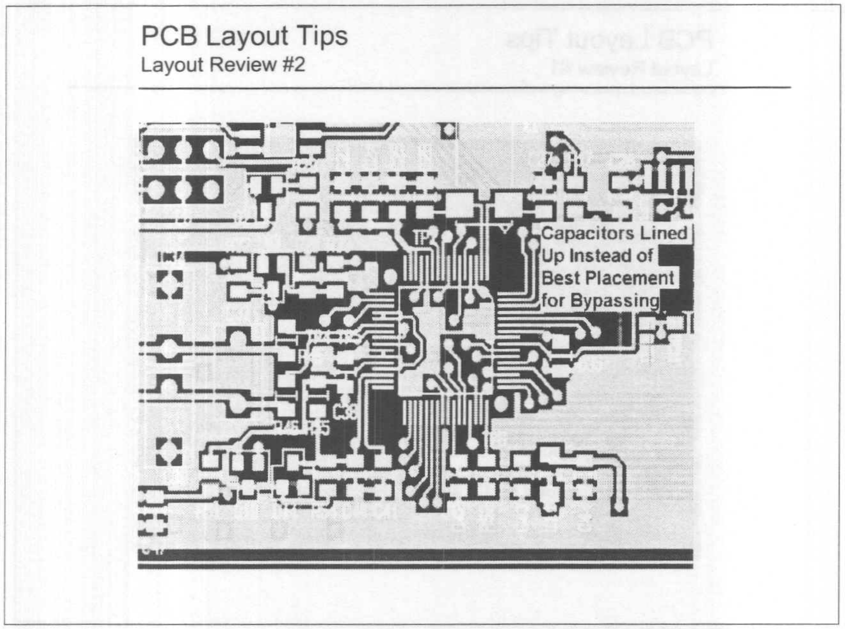


PCB Layout Tips

Layout Review #1

Ground Current from U3 affects ground for C1 & C2 which affects Crystal Y1

The ground current from U3 was routed to the ground point for the crystal (Y1) capacitors and then to the rest of the ground in the system. This caused a significant effect on the oscillator and made the clock very unstable.

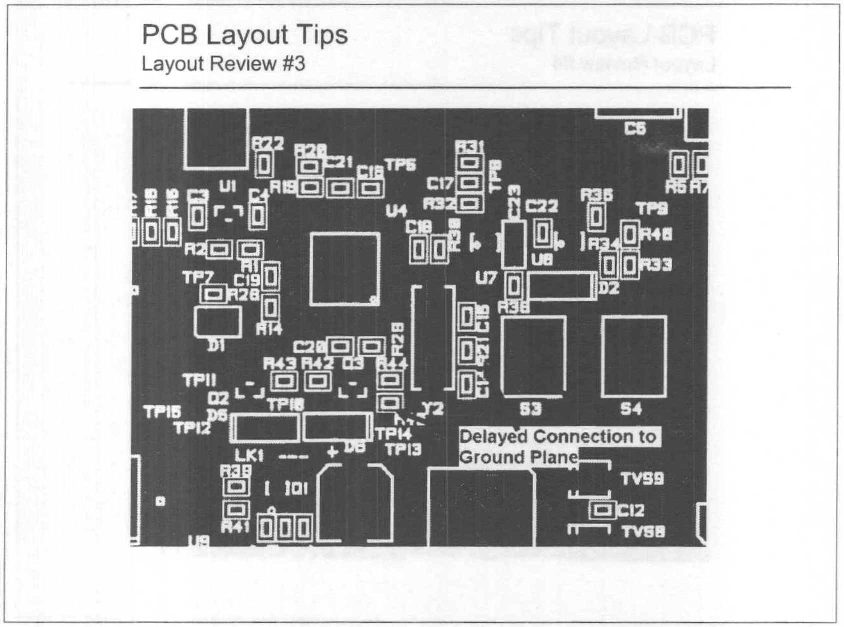


PCB Layout Tips

Layout Review #2

Capacitors Lined Up Instead of Best Placement for Bypassing

If we position capacitors and components with an order and goal of symmetry, the results could be unnecessary inductance and resistance that are introduced which will reduce the performance of the system.



PCB Layout Tips

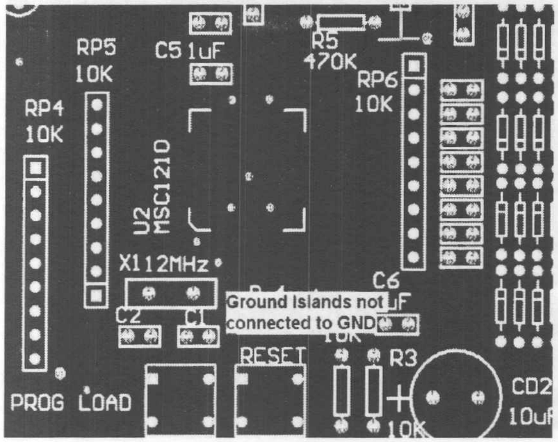
Layout Review #3

Delayed Connection to Ground Plane

Here this board is designed with a reasonably solid ground plane and rather than connecting directly with feedthroughs to ground plane, there are large traces to a centralized ground connection point. The extra inductance and resistance, even with wide traces, will reduce the performance of the board compared to feedthroughs (or multiple feedthroughs) placed close to the component's ground connection.

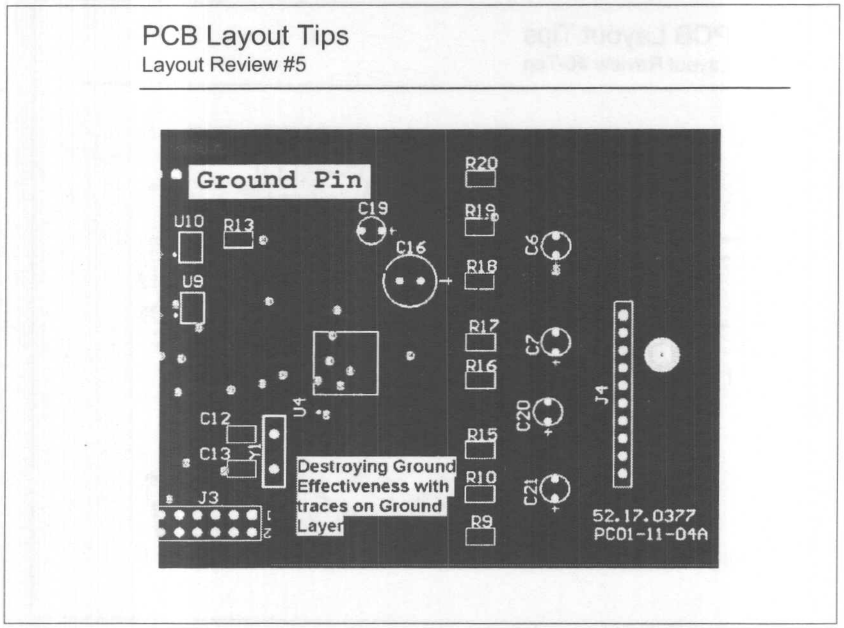
PCB Layout Tips

Layout Review #4



Ground Islands not connected to GND

These top planes seem to be just an area fill for the gaps in the traces. When the plane areas on the top layer are not connected to the ground plane on the bottom, that means that the ground plane on the bottom doesn't have a continuous path available for current. If these top ground planes were connected to the bottom ground plane, the ground return currents would have a lower inductance path.



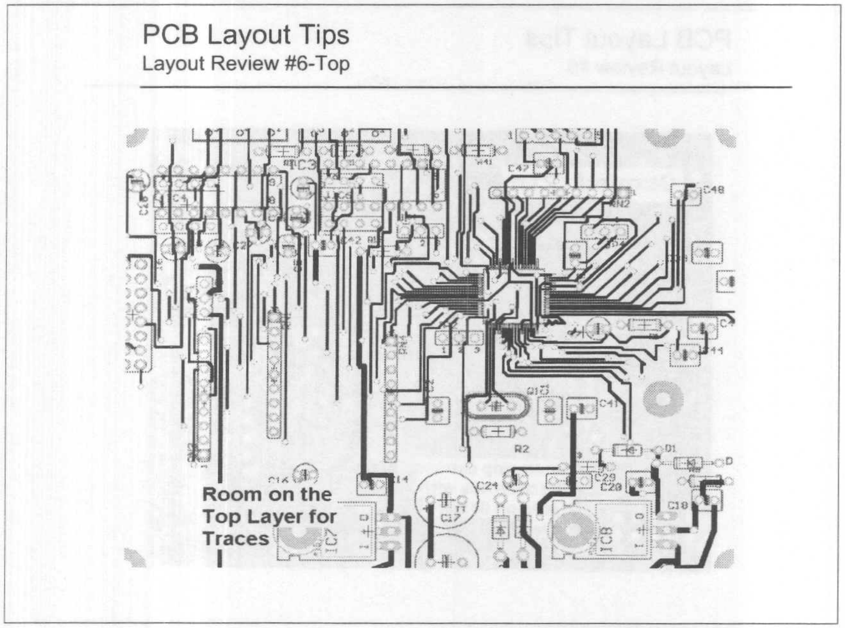
PCB Layout Tips

Layout Review #5

Ground Pin

Destroying Ground Effectiveness with traces on Ground Layer

Here we see serious splits in the bottom ground plane. A close examination will also show that the traces that are splitting the ground plane are actually also ground signals.

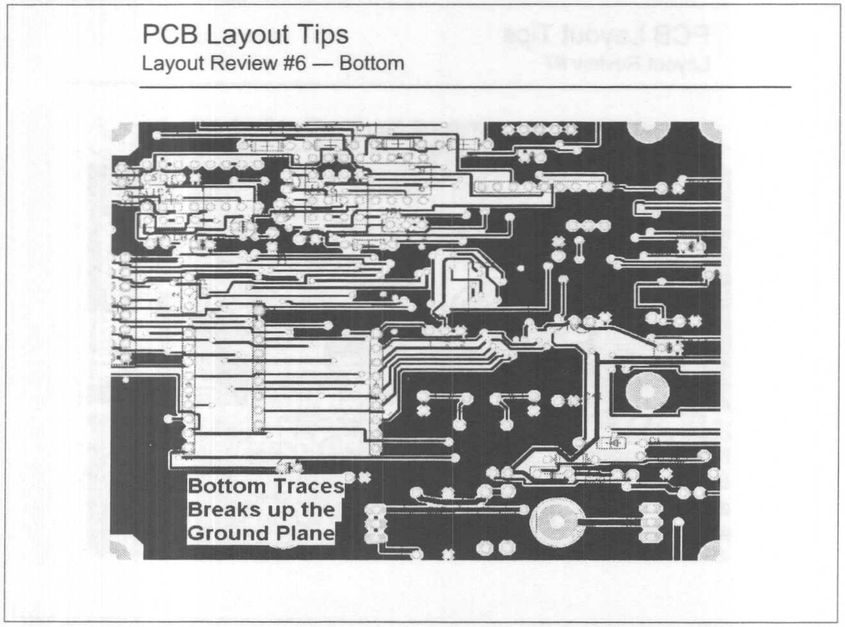


PCB Layout Tips

Layout Review #6-Top

Room on the Top Layer for Traces

As this layer is compared with the next slide, it is evident that some of the traces could have been put on this layer so that the ground layer could have been more solid.

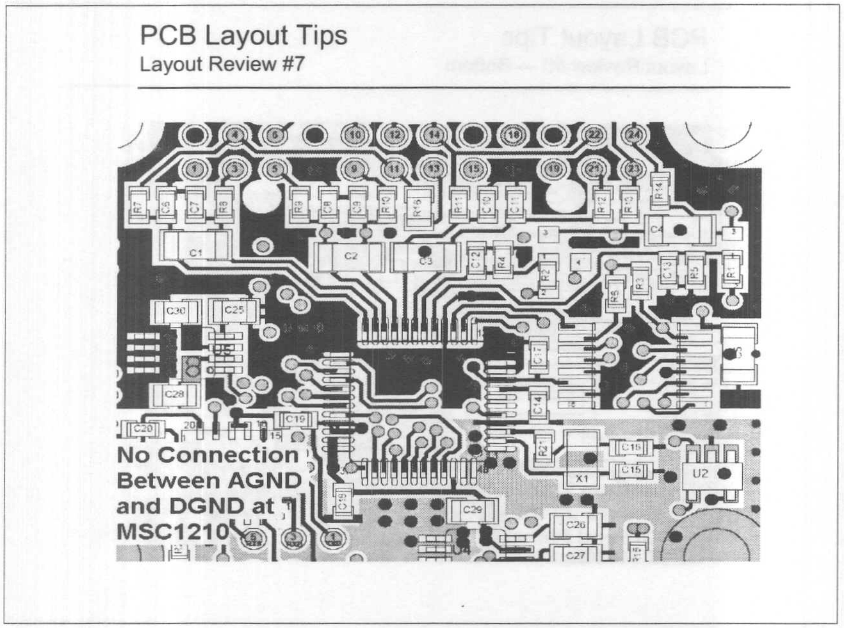


PCB Layout Tips

Layout Review #6 — Bottom

Bottom Traces Breaks up the Ground Plane

This shows the problem of using a digital PCB layout for an analog board. It simplifies the routing for most of the traces on each layer to go one direction, but that might lead to inefficiencies in the ground return currents and an unnecessary increase in inductance. Remember, inductance is defined by the size of the loop for the current.

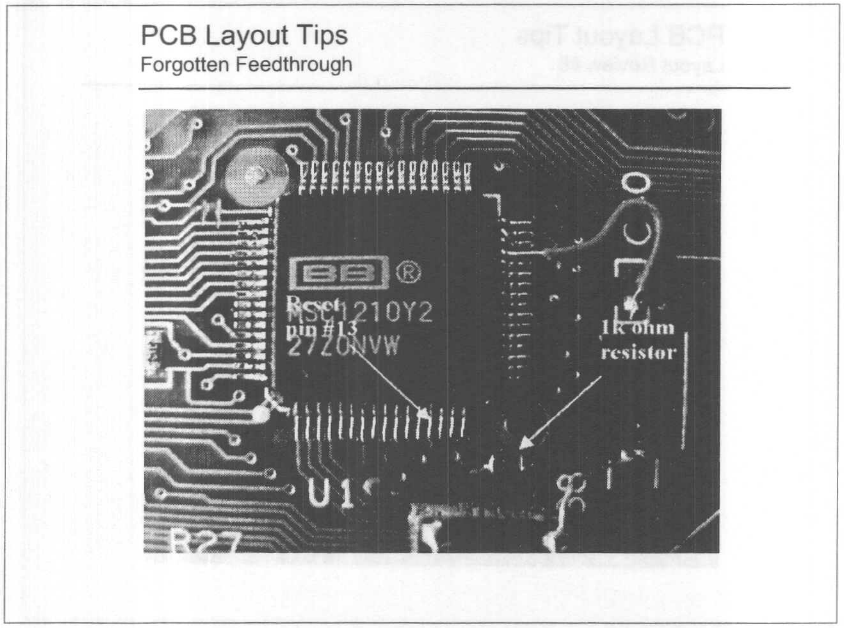


PCB Layout Tips

Layout Review #7

No Connection Between AGND and DGND at MSC1210

There is a good separation between the analog and digital sections, but the AGND and DGND should be connected together at the ADC.



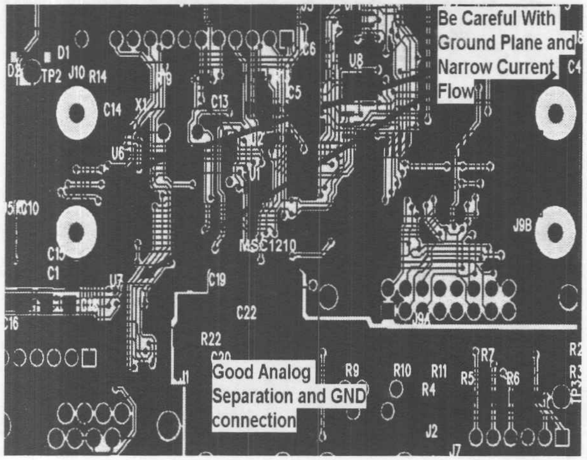
PCB Layout Tips

Forgotten Feedthrough

The goal was to place the resistor to protect the reset pin. The close placement of the resistor didn't solve the problem. However the feedthrough was to an external connector that defeated the purpose of the resistor.

PCB Layout Tips

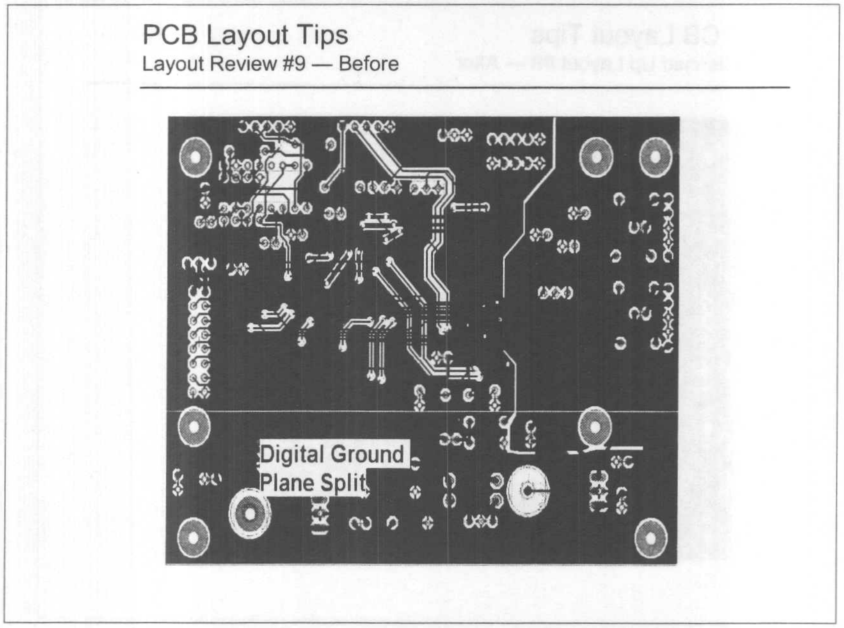
Layout Review #8



Be Careful With Ground Plane and Narrow Current Flow

Good Analog Separation and GND connection

The Analog and Digital sections are nicely separated. The AGND and DGND are connected with a wide connection right at the MSC1210. But the ground plane sometimes has narrow channels for current which will increase inductance and noise. The return currents for the digital signals will want to travel under the horizontal traces, but the channels in the ground plane are generally vertical.



PCB Layout Tips

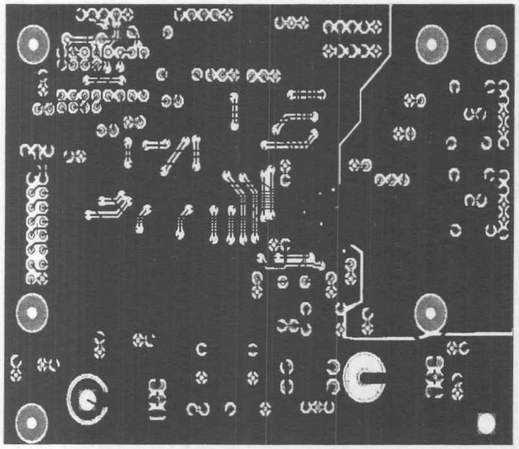
Layout Review #9 — Before

Digital Ground Plane Split

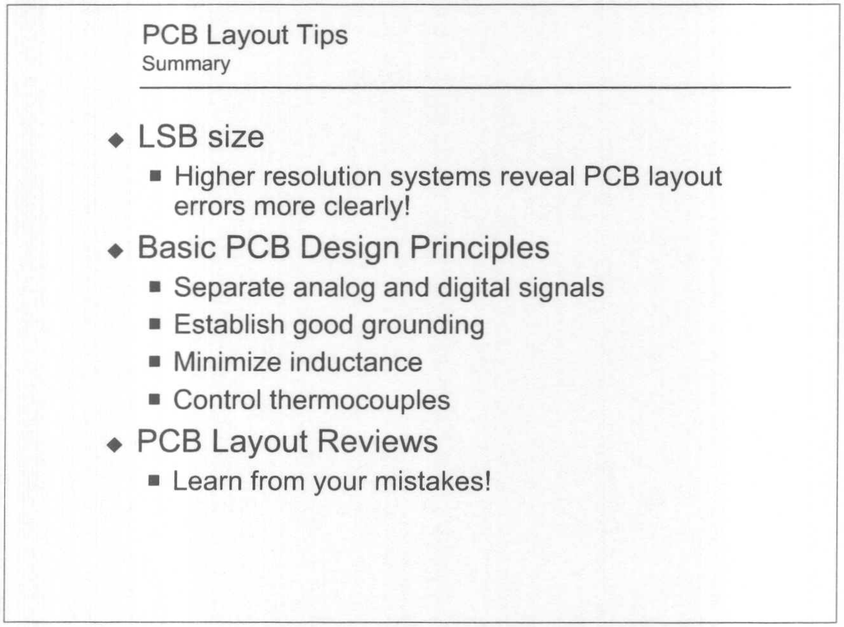
Analog and Digital sections separated with ground planes on both the bottom and the top. But the bottom ground plane is seriously split so that the ground return currents can't flow close to the source signal wires.

PCB Layout Tips

Cleaned Up Layout #9 — After



A few simple changes in the routing were able to provide a much better ground return path. Larger continuous areas of ground plane translate into short return current paths and lower inductance.



PCB Layout Tips

Summary

- LSB size
    - Higher resolution systems reveal PCB layout errors more clearly!
- Basic PCB Design Principles
    - Separate analog and digital signals
    - Establish good grounding
    - Minimize inductance
    - Control thermocouples
- PCB Layout Reviews
    - Learn from your mistakes!

Section 10

Additional Information on Products, Tools and Support

From

Texas Instruments

For more information on Data Converters please look at the Texas Instruments data converter home page:

www.dataconverter.com

This page provides access to:

DATA CONVERTER PRODUCTS

- Alphabetical Product Listing
- Device Locator
- New Products
- Parametric Search
- Part Number and Keyword Search

DESIGN RESOURCES

- Application Notes
- Datasheets
- Development Tools (EVMs)
- Packaging Information

HOW TO PURCHASE

- Distributors
- Pricing and Availability
- Samples

SUPPORT

- Ask an Expert
- Industry Forums
- News and Publications
- Standards Bodies
- Training

For more information on Amplifiers please look at the Texas Instruments amplifier home page:

www.amplifier.ti.com

This page provides access to:

AMPLIFIER PRODUCTS

- Alphabetical Product Listing
- Analog Cross Refefence Search
- Device Locator
- Parametric Search
- Part Number and Keyword Search

DESIGN RESOURCES

- Application Notes
- Datasheets
- Development Tools
- Engineering Design Utilities
- Packaging Information
- Macro Models

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INTERFACE PRODUCTS

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